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A QUANTITATIVE STUDY OF ASYMMETRIC-SIDEBAND BROADCASTING*

By P. P. ECKERSLEY, Member.

(*Paper first received* 16*th October*, 1937, *in revised form* 19*th January, and in final form* 8*th June*, 1938; *read before the* WIRELESS SECTION 2*nd March*, 1938.)

SUMMARY

The major factor which hampers the progress of broadcasting is the lack of a sufficient number of channels in which to contain the increasing number of transmissions. In the present state of the art few, except local, stations can be received without interference from other broadcasting stations on frequency-contiguous channels unless the upper frequencies of modulation of the wanted station are severely attenuated and unless, therefore, reproduction is lacking in intelligibility and naturalness. This inter-station interference can be eradicated or minimized, and quality improved as regards "top" reproduction, if all or part of one sideband of the transmitted spectrum is cut away. The cutting-away operation on the side-band, however, produces harmonic distortion in the received signals. It is the object of the asymmetric-sideband system of transmission to cut away part of one sideband without introducing audible harmonic distortion.

The paper gives a quantitative analysis which shows how asymmetry, either of phase or of magnitude, between two sideband components, created by any given modulation component, is related to harmonic distortion; and it is shown that this distortion is proportional, over a large range of values, to depth of modulation. The nature of the sound spectrum is such that the intensities in the upper register are much less than those in the lower middle register. Since the degree of asymmetry between counterpart sideband components is a measure of the attenuation of the cut-away sideband, this attenuation can be increased as the modulation frequency is made greater. This is tantamount to saying that, as the sideband frequency becomes more and more removed from the carrier frequency, it may be more and more attenuated while the distortion can be made to remain constant and small. Thus in asymmetric technique the outer parts of one sideband may be removed without the introduction of audible distortion.

Curves are derived, based upon the evaluation of expressions given in the quantitative analysis, which show the required attenuation-constant of filters which cut away part of one sideband but produce no more than a constant distortion. It is proved that phase asymmetry between counterpart sidebands may give rise to more serious distortions than are created by magnitude asymmetry acting alone, and a network is described which keeps phase asymmetry to a minimum and yet gives a close approximation to the required attenuation.

Accounts are given of practical tests wherein the quality of reproduction obtainable from an asymmetric transmitter is compared with that given by orthodox modulation. While it is true that a highly trained ear may, on rare occasions, be able to detect some slight differences between the two types of reproduction, when they are directly compared, the general public would be quite incapable of so doing, even if they possessed receiving apparatus so free from distortion itself as to give them an opportunity to do so. It is pointed out that it is better to face the occasional minute deteriorations in quality given by the asymmetric system than to suffer from the continuous hissing and splitting sounds brought about by existing inter-station interference.

In the latter part of the paper an analysis is given which shows that the asymmetric system has three possible applications to space broadcasting as it is practised to-day, namely (1) It could be used to allow carrier difference frequencies of the order of 11–12 kc. without introducing any sideband overlap interference. (2) It could be applied to existing conditions, where carrier difference frequencies are of the order of 9 kc., and would reduce inter-station interference to the order of one-tenth its present value. (3) If existing interferences are considered to be tolerable, then carrier difference frequencies can be reduced to 6 kc. without increasing interference above its existing values; this implies that 1½ times the number of stations working in Europe to-day on exclusive waves could be accommodated in the existing waveband.

It is suggested, finally, that a policy could be adopted such that the introduction of the asymmetric system would encourage receiver development along lines wherein, by an augmentation of only the carrier component at the detector, all distortion due to the use of an asymmetric spectrum would be eliminated. Thus an evolution might be envisaged which led to the introduction of the ideal carrier and single-sideband type of transmission when all receivers were adapted to reduce the resulting distortion to a tolerable amount.

LIST OF SYMBOLS

$a = \frac{M_N}{2}(1 + \xi)$.

$b = \frac{M_N}{2}(1 - \xi)$.

$d = 4\delta/M_N$, d having suffixes to distinguish the nature of δ.

f_m = mean pass frequency of band-pass filter.

f_s = frequency of sideband component.

f_0 = carrier frequency.

f_1 = lower frequency of cut-off in band-pass filter.

f_2 = upper frequency of cut-off in band-pass filter.

Δf = frequency of modulating component.

Δf_∞ = a modulating frequency which, added to or subtracted from the carrier frequency, expresses a frequency at which α may be infinite.

M_A = modulation factor of carrier after passing through asymmetric filter.

$M_{A\xi}$ = modulation factor of carrier after passing through asymmetric filter when only magnitude asymmetry exists.

$M_{A\phi}$ = modulation factor of carrier after passing through asymmetric filter when only phase asymmetry exists.

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* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 36.

LIST OF SYMBOLS—*continued*

M_N = modulation factor of carrier before passing through asymmetric filter.
S = magnitude of one sideband-component vector.
t = time.
α_ϵ = attenuation constant of equalizer.
α_R = desirable attenuation constant of asymmetric carrier-frequency filter if no demodulation correction takes place.
α_T = desirable attenuation constant of asymmetric carrier-frequency filter with demodulation correction.
δ = distortion factor, i.e. ratio of second-harmonic amplitude to fundamental amplitude.
δ_ξ = distortion factor produced by magnitude asymmetry alone.
δ_ϕ = distortion factor produced by phase asymmetry alone.
δ_0 = constant value of distortion chosen as tolerable or negligible in asymmetric systems.
θ_1 and θ_2 = phase angles made by two sideband components, created by the same modulating component, with carrier component.
ξ = ratio (< 1) of amplitudes of two sideband components created by the same modulating component.
2ϕ = algebraic sum of θ_1 and θ_2.
ψ = angle between vector resultant of two unequal-magnitude sideband vectors and the vertical.
$\Delta\omega = 2\pi\Delta f$.

(1) INTRODUCTION

A paper outlining the principles of the asymmetric-sideband system has already been published.* That paper, which was designed to be no more than a preliminary survey of principle, is supplemented herein by a quantitative analysis which forms a guide to practical design.

The object of the asymmetric-sideband system of transmission is to reduce the frequency band width normally occupied when a carrier is modulated (and when both sidebands are fully transmitted) without the introduction of audible distortion into the received signals. If this aim could be practically achieved, then the practice of broadcasting could be benefited in any or some of the following ways:—

(*a*) The existing carrier difference frequencies could be slightly increased and all sideband overlap interference would thereby be eliminated, a benefit *per se*, and at the same time the quality of reproduction of radio programmes with respect to the reproduction of the higher audio frequencies could be greatly improved.

(*b*) The interferences associated, in the modern practice of broadcasting, with the working of stations at carrier difference frequencies less than twice the highest modulation frequency transmitted, would be reduced or eliminated.

(*c*) If the interferences referred to in (*b*) are considered to be, in all the circumstances, tolerable, then, if the asymmetric system was used, existing carrier difference frequencies could be reduced, and the number of stations which could work simultaneously in a given continental area could be thereby increased.

One of the most important features of the system is that no alteration to existing receivers would be immediately necessary when the transmitters were modified to radiate the asymmetric spectrum; so far as the ordinary listener is concerned the service from local stations would appear to continue as before the changes were made, but some amelioration of inter-station interference would take place. Some modifications in reception technique would eventually be desirable, however, if the listener were to benefit to the greatest possible extent from the improved system of transmission.

(2) OUTLINE OF PRINCIPLE

It is proposed first to outline the basic principles of the asymmetric system qualitatively and thereafter to give a quantitative analysis forming a guide to practical design.

When a carrier is modulated by a lower frequency three component frequencies are created, namely the carrier, an upper-frequency sideband component, and a lower-frequency sideband component. The asymmetric system is based upon the use of a filter which gives, at the higher modulation frequencies, a differential attenuation between the upper- and lower-frequency sideband components created by a given modulation component. This filter or a series of filters cutting away part of one sideband is connected at any point in the chain of circuits between the output of the circuit where the normally modulated carrier is generated and the input to the receiver detector. When the asymmetric system is applied, as for its greatest usefulness it should be, to the transmitter, a sideband-cutting filter is connected at any point in the chain of circuits between the terminals where the normally modulated carrier is generated and the input to the transmitting aerial.

The filter used for cutting away part of one sideband is designed to transmit the two sideband components created by low frequencies of modulation (and therefore having frequencies close to the carrier frequency) at substantially equal amplitude but to increase the attenuation of one sideband component relatively to its fully transmitted counterpart as the frequency of modulation (and hence the frequency difference of the sideband components from the carrier frequency) becomes greater. Fig. 1 illustrates the principle of the system. The Figure shows the frequency/attenuation and frequency/phase characteristics of a band filter. f_0 is the carrier frequency, located asymmetrically with respect to the mean pass frequency of the band filter f_m. $(f_0 \pm \Delta f_1)$ and $(f_0 \pm \Delta f_2)$ are sideband component frequencies corresponding to modulation frequencies Δf_1 and Δf_2 respectively. When the modulation frequency is relatively low and has a value Δf_1 both sideband components are substantially equally transmitted, but when the modulation frequency has a value Δf_2, greater than Δf_1, one sideband component, of frequency $(f_0 - \Delta f_2)$, is attenuated, while the other, of frequency $(f_0 + \Delta f_2)$, is fully transmitted.

Asymmetry between two counterpart sideband components introduces harmonic distortion in the received signals. This asymmetry may either be an asymmetry of magnitude or the sideband components may have a phase asymmetry to the carrier. Both forms of asymmetry may be present simultaneously, both produce harmonic distortion. It will be shown in the quantitative analysis that the distortion introduced is, in practical conditions of working, generally proportional to modulation depth and is a function of the degree of asymmetry.

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* *Journal I.E.E.*, 1935, vol. 77, p. 517; and *Proceedings of the Wireless Section*, 1935, vol. 10, p. 177.

It is known that the intensity of the components in the spectrum of sounds forming typical speech and music radio programmes falls rapidly as the frequency of these components increases above a certain frequency (of the order of 1 000–1 500 cycles per sec.). Hence if some value of distortion can be regarded, thanks to the ear's tolerance, as negligible, this distortion may be kept constant (and negligible) in an asymmetric-sideband system provided the degree of asymmetry between counterpart sideband components is made a function of the probable depth of modulation occurring in practice. The fact that modulation depth decreases as modulation frequency increases makes it possible to cut away the components of one sideband more and more as they are of greater frequency-difference from the carrier without introducing more than a constant and negligible distortion. Thus in the asymmetric system the bulk of the modulated energy is radiated in a symmetric form; asymmetry is only introduced in frequency bands where the sideband energy is relatively small. Since in the proposed adaptation of orthodox technique a double-sideband system carries the bulk of the energy, and because the distortion due to the cutting-away of the remaining spectrum is so slight, it is obvious that the average receiver will not remark any difference in harmonic distortion between the symmetric and asymmetric transmissions.

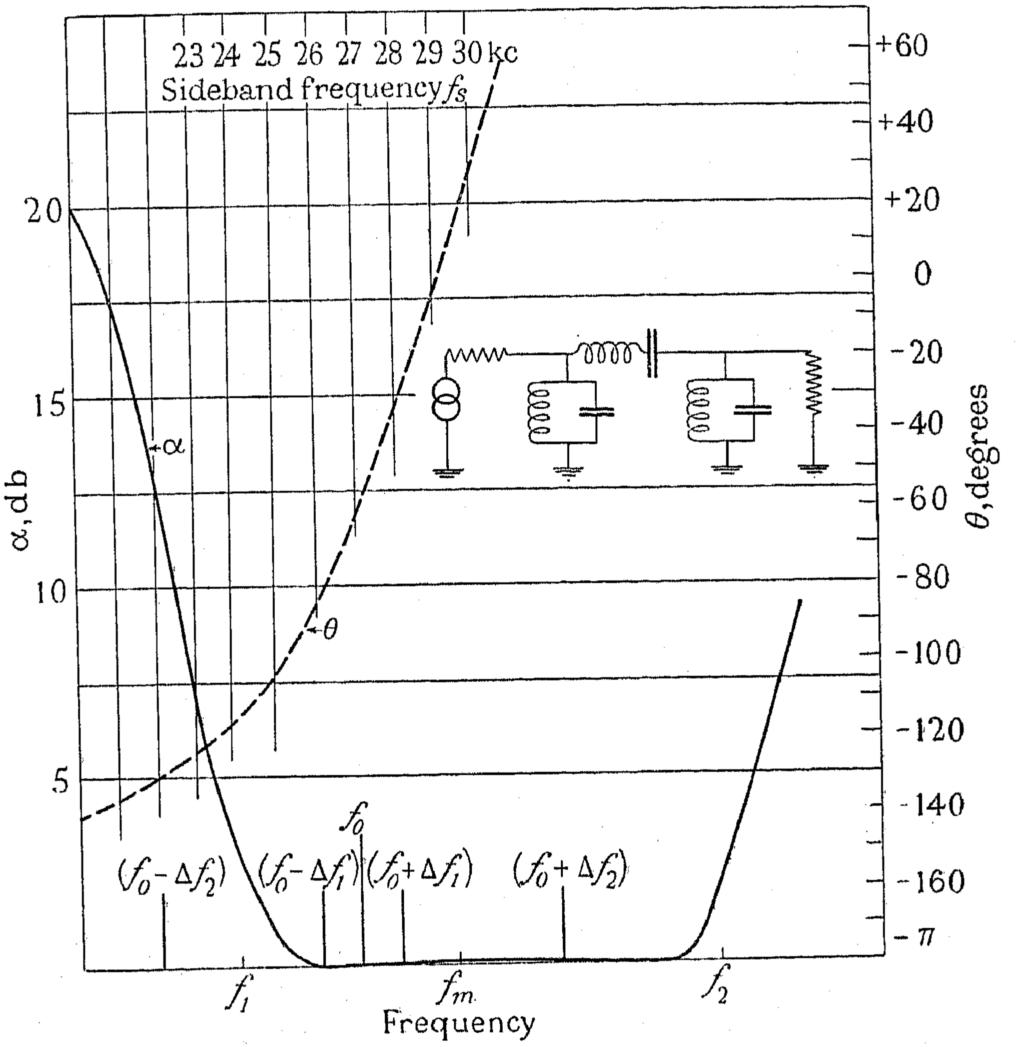

Fig. 1

Points obtained experimentally. $f_0 = 27$ kc./sec., $\Delta f_1 = 1$ kc./sec., $\Delta f_2 = 5$ kc./sec.

The differential treatment of the sidebands, however, introduces not only harmonic distortion but, unless precautions are taken to correct it, frequency-characteristic distortion as well. By the frequency characteristic is meant the curve showing the intensity of the modulated output plotted against frequency, for a constant input modulation. Let M_N be the modulation factor of the carrier before passing through the asymmetric filter and let M_A be the modulation factor of the carrier after passing through the asymmetric filter. Suppose, further, that the modulation frequency be varied while M_N is kept constant. At low frequencies of modulation both sideband components are equally transmitted by the filter, and $M_A = M_N$. At a frequency of modulation so high that one sideband component is relatively of zero amplitude compared with the other, the effective modulation of the carrier at the filter output terminals will be half its value at the input to the filter, or, for high modulation frequencies, $M_A = \frac{1}{2}M_N$. Thus the overall frequency-characteristic of an "uncorrected" asymmetric-sideband transmitter would be represented by a curve showing a maximum at low modulation frequencies and falling through intermediate values to half that maximum at high modulation frequencies. Thus the action of the sideband-cutting filter is partly to demodulate the normally modulated carrier. The distortion of the frequency characteristic may be corrected by connecting somewhere, between the input modulation terminals and the input to the modulating valve, a frequency-discriminating network designed to attenuate the modulations by half for the lower frequencies but to give them full transmission at high frequencies.

It is necessary that the depth of modulation of the carrier in an asymmetric transmitter shall be sensibly the same as that obtainable in normal practice. Thus the asymmetric transmitter must embody, as well as the equalizer which attenuates the lower frequencies of the modulating spectrum, extra amplification, above that necessary for conventional technique, to raise the level of the modulations by an amount equal to the maximum equalizer attenuation.

The result of amplification and attenuation in the correction system is to leave the intensity of the lower-frequency modulating components the same as if no corrector circuits had been interposed. The higher-frequency components of modulation are more and more amplified as their frequency is higher, the maximum occurring when the equalizer is fully transmitting and when the amplification doubles the normal intensity. The amplitude of any higher frequency can never exceed the normal maximum because the design of the equalizer circuits is proportioned by the audio spectrum intensities and so no overload takes place albeit the apparent modulation in each system, asymmetric and normal, is the same.* It is thus seen that, provided the circuits are correctly designed, the asymmetric system need introduce neither serious harmonic distortion nor any frequency-characteristic distortion even though a part of the normally radiated spectrum is cut away.

The root question, the answer to which will alone determine whether or not the scheme has practical value, is whether the amount by which the normal spectrum may be narrowed is sufficient to justify the introduction of the resulting harmonic distortion. Evidently the quantities involved might be such that, in order to keep the distortion negligible, it would not be possible to cut away more than a small outer part of one sideband. On the other hand, a study might reveal that one sideband could be almost wholly eliminated without the harmonic distortion becoming intolerable. The question at issue can obviously only be answered in terms of a quantitative analysis, and it is therefore necessary to determine the relationship between the relevant parameters. More specifically, a quantitative analysis must give:—

(*a*) The relationship between harmonic distortion, modulation factor, and the degree of asymmetry between counterpart sideband components.

(*b*) The relationships between frequency-characteristic distortion and the degree of asymmetry between counterpart sideband components.

(*c*) The relationship between modulation frequency and modulation factor in the spectrum of sounds forming typical broadcast intelligence, i.e. the frequency/intensity relationship in the spectrum of normal sounds transmitted in radio broadcasting.

(3) QUANTITATIVE ANALYSIS

General

Fig. 1, already referred to, shows a typical attenuation and phase characteristic of a single-section constant-k type of band filter working between resistance terminations which match, or tend to match, the filter image impedance. f_1 and f_2 are the cut-off frequencies, and f_m = mean pass frequency = $\sqrt{(f_1 f_2)}$. In the asymmetric system the carrier frequency is located between f_1 and f_m but closer to f_1, or between f_2 and f_m but closer to f_2. If, in Fig. 1, the frequency of modulation is Δf_2, then the amplitude of one sideband component of frequency $(f_0 - \Delta f_2)$ is less than the amplitude of the counterpart sideband component of frequency $(f_0 + \Delta f_2)$. But if the frequency of modulation is comparatively low and equal to Δf_1 then both the sideband components are substantially equally transmitted.

If the ratio of amplitudes of counterpart sideband components created by a given frequency of modulation is ξ, a ratio less than unity, then the magnitude of one sideband component (in the case considered that of lower frequency) is ξ times that of the other. We may define ξ as the magnitude asymmetry factor, the value of which may vary between zero and unity.

A consideration of the phase characteristic of the filter will show that if the carrier is located near one cut-off frequency then the phase angle of one sideband component to the carrier is different from the phase angle of the other sideband component to the carrier. If θ_1 and θ_2 be these angles, then the algebraic sum $(\theta_1 + \theta_2)$ may be finite. We shall, for reasons which will be clear later, write $(\theta_1 + \theta_2)/2 = \phi$, and call ϕ the phase-asymmetry factor. If the carrier has a frequency f_m, ϕ is always zero and phase asymmetry is wholly absent because only at this frequency is one half of the phase characteristic of the filter a mirror image of the other half.

It will be helpful in developing the analysis to consider special cases, as follows:—

(i) $\xi = 1, \phi = 0$ (symmetric and normal transmission);
(ii) $\xi < 1, \phi = 0$ (asymmetric transmission with amplitude asymmetry only);
(iii) $\xi = 1$, ϕ finite (phase asymmetry only);
(iv) $\xi < 1$, ϕ finite (asymmetric transmission with both amplitude and phase asymmetry, i.e. the general case).

Harmonic distortion when $\xi = 1$, $\phi = 0$ (symmetric transmission, Fig. 2)

It is convenient, both to illustrate the method of the more complete analysis and to define symbols, to recapitulate a well-known analysis of modulation.

It is the object throughout this Section to derive expressions which define the shape of the wave envelope resulting from the modulation of a high or carrier frequency by a lower frequency. The shape of the wave envelope is given (sufficiently accurately, even when distortions are present) by a curve passing through points the ordinates of which are the positive (or negative) peak amplitudes of the modulated carrier while the abscissae represent the times at which these peak amplitudes occur. It is assumed throughout the work that the modulation is sinusoidal, that the unmodulated-carrier vector amplitude is unity, and that rectification is " linear."

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* See the Section on " Equalizer Design " (page 153).

Fig. 2 shows the relevant vector diagram. OF, the unit-magnitude carrier vector, is considered to remain stationary and vertical. The sideband vectors FA and FB, each of magnitude S, are supposed to rotate in opposite directions at the same angular velocity, given by $\Delta\omega t = 2\pi\Delta f t$, where Δf is the modulation frequency and t is the time. An auxiliary circle is constructed with centre F and radius FR′ equal to the arithmetic sum $2S$ of the sideband-vector amplitudes. It is evident from a consideration of the triangle FR′R that $FR = 2S\cos\Delta\omega t$ and that the vector sum of the sideband vectors and the unit-magnitude carrier vector is $OR = 1 + 2S\cos\Delta\omega t$. If the magnitude of OR is plotted against $\Delta\omega t$ the resulting curve is a cosine curve. The peak and minimum values of the resultant are $(1 + 2S)$ and $(1 - 2S)$ respectively, corresponding to values of $\Delta\omega t$ of 0 and 180° respectively.

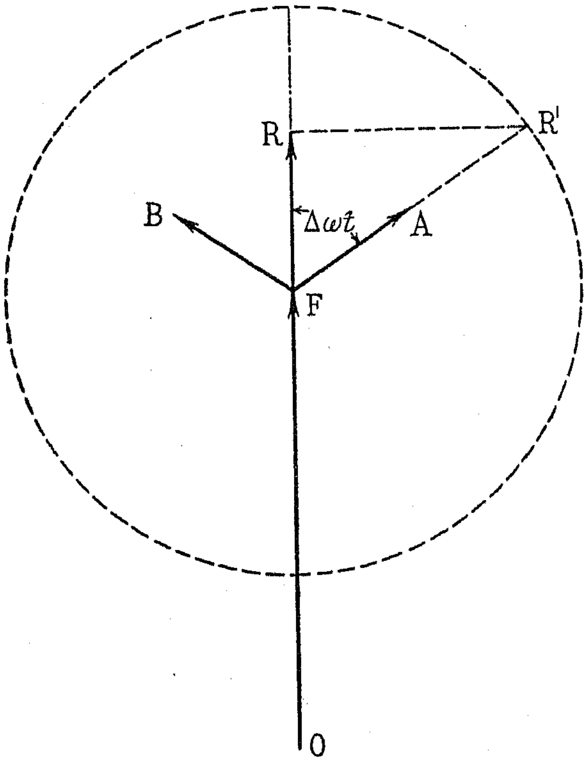

Fig. 2.—Symmetric transmission.

$\xi = 1$, $\phi = 0$, $M_N = 0{\cdot}6$.

—-—-— Locus of R.

- - - - - Construction lines.

——— Vectors.

The process of rectification eliminates the carrier; and the resulting audio component, if there are no losses, and if the rectifier is linear, is given by the function $1 + 2S\cos\Delta\omega t - 1 = 2S\cos\Delta\omega t$. Thus the audio-component intensity is represented by the wave-envelope curve if this is drawn about a mean line whose height, above the horizontal zero axis passing through 0 (Fig. 2), is equal to the carrier amplitude.

The modulation constant is defined as the ratio of the incremental peak volts, when modulation takes place, to the value of the carrier peak volts, or

$$M_N = \frac{2S}{1} = 2S \quad . \quad . \quad . \quad . \quad . \quad (1)$$

Thus the values of the audio components after rectification are proportional to M_N and there is no distortion.

Harmonic distortion when $\xi < 1$, $\phi = 0$ (asymmetric transmission with only magnitude asymmetry, Fig. 3)

Fig. 3 is identical with Fig. 2 except that the vector FB is ξ times FA, ξ being the magnitude-asymmetry factor and having any value between unity and zero. An auxiliary circle is drawn of radius FR′ which is the arithmetic sum of FB and FA. A point G is chosen on a horizontal line passing through F so that FG is the arithmetic difference between FA and FB. FR (not drawn) is the resultant of FA and FB. Knowing that $FB = \xi FA$, that $FA = S$, and that $S = M_N/2$ gives the equivalences

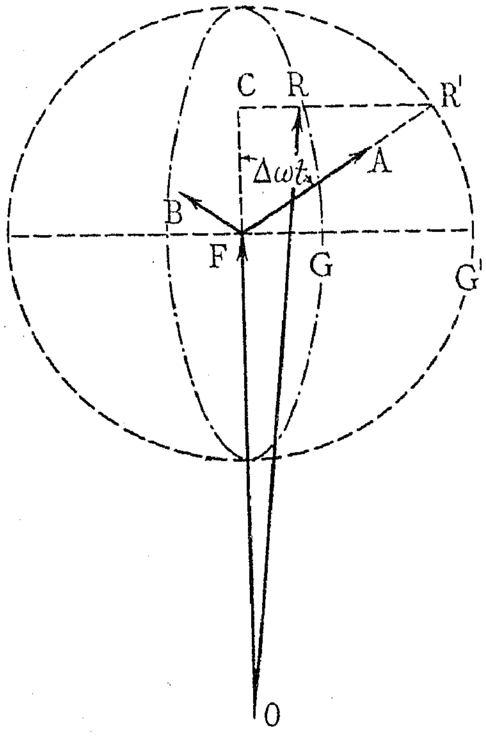

Fig. 3.—Asymmetric transmission, magnitude asymmetry only.

$\xi = 0{\cdot}5$, $\phi = 0$, $M_N = 0{\cdot}6$, $M_{A\xi} = 0{\cdot}45$.

—-—-— Locus of R.

- - - - - Construction lines.

——— Vectors.

$$FR' = a \text{ and } FG = b$$

where
$$a = \frac{M_N}{2}(1 + \xi) \quad . \quad . \quad . \quad . \quad . \quad (2)$$

and
$$b = \frac{M_N}{2}(1 - \xi) \quad . \quad . \quad . \quad . \quad . \quad (3)$$

The resultant of the three vectors OF (the carrier vector) and FB and FA (the sideband component vectors) is OR. The value of OR may be found at any instant of time by finding expressions for the x and y co-ordinates, namely CR and (OF + FC), of the point R.

It can be shown that

$$x = b\sin\Delta\omega t \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

and
$$y = 1 + a\cos\Delta\omega t \quad . \quad . \quad . \quad . \quad (5)$$

whence

$$OR = (a^2\cos^2\Delta\omega t + b^2\sin^2\Delta\omega t + 2a\cos\Delta\omega t + 1)^{\frac{1}{2}} \quad (6)$$

It would be possible to derive the fundamental and harmonic values from expression (6). It is more logical, however, to derive a comprehensive expression dealing with the general case where ξ is less than unity and ϕ is finite, when the value of the distortion component can be found for particular cases by writing $\phi = 0$ and $\xi < 1$, or $\xi = 1$ and ϕ finite, in such a general expression. In order, however, to check the findings of the analysis, Appendix I gives a simple method for finding an expression for the distortion when $\phi = 0$ and $\xi < 1$. In the meantime a further analysis of the case here dealt with may help to clarify what follows.

The locus of R is an ellipse having axes given by the vector magnitude of FR. Evidently, from (4) and (5),

$$\mathrm{FR} = (b^2 \sin^2 \Delta\omega t + a^2 \cos^2 \Delta\omega t)^{\frac{1}{2}}$$

or, substituting the values of a and b in terms of M_N and ξ,

$$\mathrm{FR} = \frac{M_N}{2}(1 + 2\xi \cos 2\Delta\omega t + \xi^2)^{\frac{1}{2}} \quad . \quad . \quad (7)$$

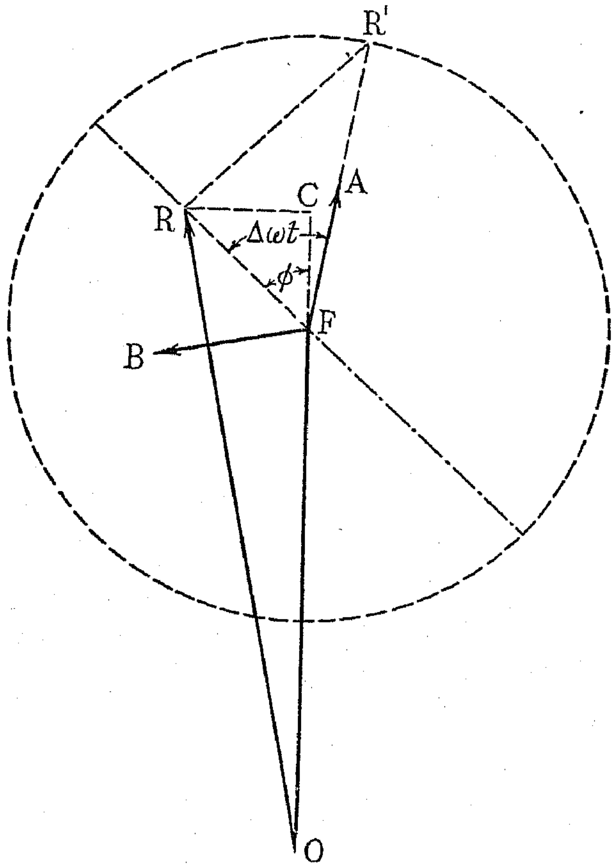

Fig. 4.—Asymmetric transmission, phase asymmetry only.
$\xi = 1$, $\phi = 45°$, $M_N = 0{\cdot}6$, $M_{A\phi} = 0{\cdot}54$.
— - — Locus of R.
- - - - Construction lines.
——— Vectors.

The angle made by FR with the vertical is

$$\psi = \Delta\omega t - \text{arc tan}\,\frac{\xi \sin 2\Delta\omega t}{1 + \xi \cos 2\Delta\omega t} \quad . \quad . \quad (8)$$

If $\Delta\omega t = 0$ in equations (7) and (8),

$$(\mathrm{FR})_{\Delta\omega t=0} = \frac{M_N}{2}(1 + \xi) = a, \text{ and } (\mathrm{FR})_{\Delta\omega t=0} \text{ is vertical.}$$

Or, if $\Delta\omega t = 90°$,

$$(\mathrm{FR})_{\Delta\omega t=90°} = \frac{M_N}{2}(1 - \xi) = b, \text{ and } (\mathrm{FR})_{\Delta\omega t=90°} \text{ is horizontal}$$

which is obvious from Fig. 3.

In the limits $\xi = 1$, $\xi = 0$, the locus of R becomes either a straight line (symmetric modulation) or a circle (single-sideband transmission). A diagram showing the locus of R and the wave-form resulting for the values $\xi = 0{\cdot}2$, $M_N = 1$, is given in Fig. 6.

The increment of peak volts above the carrier due to modulation, divided by the carrier amplitude, is called the modulation factor. Asymmetry between counterpart sidebands introduces a demodulating effect, and the modulation factor no longer has a value M_N but a new and smaller value M_A.

From the expressions derived, and by a consideration of Figs. 3 and 6, it can be seen that we may write

$$M_A = a = \frac{M_N}{2}(1 + \xi) \quad . \quad . \quad . \quad . \quad (9)$$

and that $M_A = M_N$ when $\xi = 1$ (symmetric modulation) and that $M_A = \frac{1}{2}M_N$ when $\xi = 0$ (single-sideband transmission).

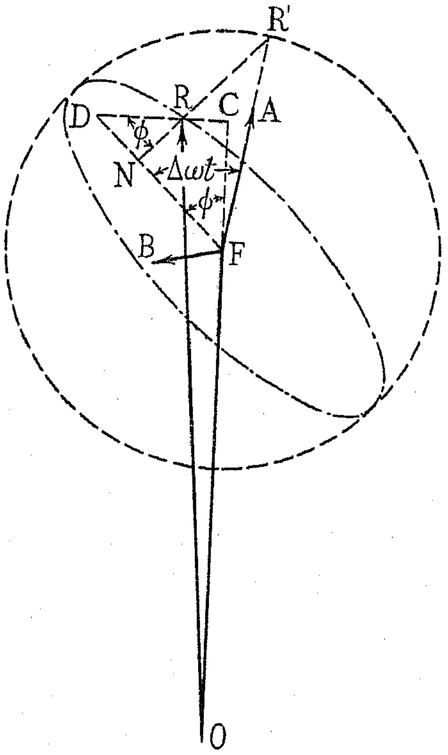

Fig. 5.—Asymmetric transmission, general case.
$\xi = 0{\cdot}5$, $\phi = 45°$, $M_N = 0{\cdot}6$, $M_A = 0{\cdot}41$.
— - — Locus of R.
- - - - Construction lines.
——— Vectors.

It is seen that the resultant vector OR swings from one side to the other of the vertical and that this swinging is a measure of the distortion. As ξ decreases so the divergence of the magnitude of the resultant from its value when $\xi = 1$ becomes greater, reaching a maximum when $\xi = 0$, i.e. when one sideband is completely removed. The divergence is a maximum when $\Delta\omega t = 90°$ and 270°, i.e. when, in symmetric modulation, the carrier has its unmodulated amplitude. Clearly the distortion is almost wholly in the nature of a 2nd harmonic. It will be assumed hereafter that the distortion factor is measured wholly by the amplitude of the 2nd harmonic divided by the fundamental.

Harmonic distortion when $\xi = 1$ and ϕ is finite (asymmetric transmission with only phase asymmetry, Fig. 4)

It is extremely unlikely, in the practice of asymmetric transmission or reception, that phase asymmetry will be considerable if the magnitude ratio of the sidebands is unity, but the case is treated separately here in order to lead up in a logical way to the general case where ξ and ϕ are both variables. As has been shown elsewhere,* when the sideband-component vectors have equal amplitude but unequal phase-angles with respect to the carrier vector the locus of their resultant (R, Fig. 4) lies along a straight line inclined to the vertical at an angle ϕ, which is equal to $(\theta_1 + \theta_2)/2$ where θ_1 and θ_2 are the angles made by the sideband-component vector with the carrier.

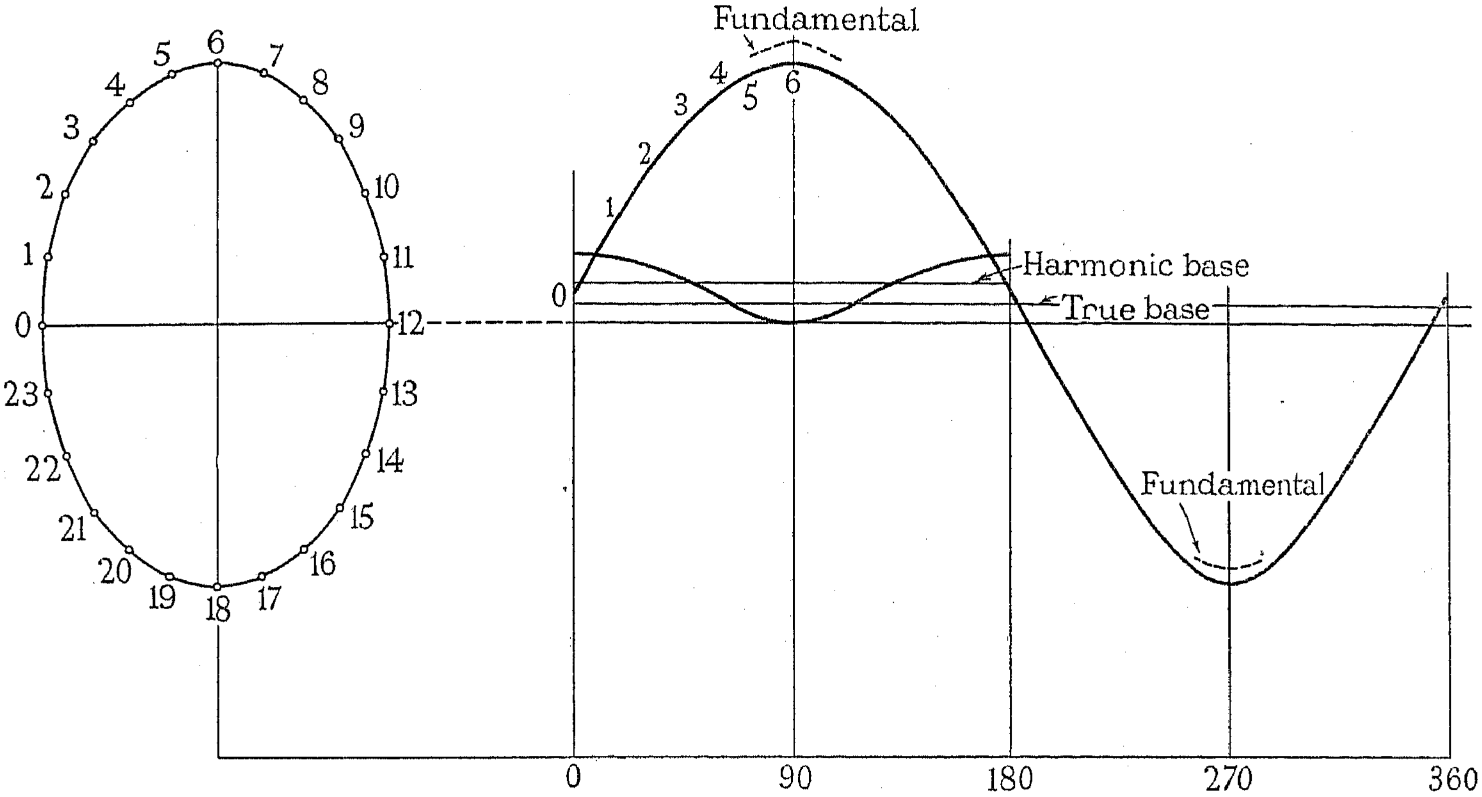

Fig. 6

$M_N = 1$, $\xi = 0{\cdot}2$, $2\phi = 0$, $M_A = 0{\cdot}6$, $\delta_\xi = 7{\cdot}0$ %.

The angle 2ϕ is in fact the algebraic sum of the sideband phase-angles and is finite when $\theta_1 \neq \theta_2$. ϕ is thus the phase-asymmetry distortion-factor.

A diagram illustrating the case is given in Fig. 4, and it can be shown from this that the co-ordinates of R are given by

$$x = M_N \cos \Delta\omega t \sin \phi \quad . \quad . \quad . \quad . \quad (10)$$

$$y = 1 + M_N \cos \Delta\omega t \cos \phi \quad . \quad . \quad . \quad (11)$$

whence

$$\text{OR} = (M_N^2 \cos^2 \Delta\omega t + 2M_N \cos \Delta\omega t \cos \phi + 1)^{\frac{1}{2}} \quad (12)$$

Evidently, by writing $\Delta\omega t = 0$ or $\Delta\omega t = 180°$ respectively,

$$(\text{OR})_{max.} = (M_N^2 + 2M_N \cos \phi + 1)^{\frac{1}{2}} \quad . \quad (13\text{A})$$

$$(\text{OR})_{min.} = (M_N^2 - 2M_N \cos \phi + 1)^{\frac{1}{2}} \quad . \quad (13\text{B})$$

If $\phi = 90°$ $(\theta_1 + \theta_2 = 180°)$,

$$(\text{OR})_{max.} = (\text{OR})_{min.} = (M_N^2 + 1)^{\frac{1}{2}}$$

and the resulting wave-envelope contains no fundamental component and is wholly of the form of a harmonic of the modulating component. This fact was revealed by Johnstone and Wright.*

A diagram showing the resulting wave-form distortion is given in Fig. 7 for certain values of M_N and ϕ. The maximum distortion is seen to occur when $\Delta\omega t = 0$ and 180°; this proves that, while phase asymmetry, like magnitude asymmetry between sideband components, produces a swinging of the resultant OR about the vertical, the resulting harmonic, which is also mainly of double the fundamental frequency, has a 90° phase-difference from that produced by magnitude asymmetry.

An approximate expression for the distortion is given in Appendix I.

Harmonic distortion when $\xi < 1$ and ϕ is finite (general case of asymmetric transmission)

From previous reasoning it is clear that when $\xi < 1$ and ϕ is finite the locus of R lies on an ellipse the major axis of which is inclined at an angle ϕ to the vertical. Three examples are illustrated in Figs. 8, 9, and 10.

It can be proved from Fig. 5 that

$$x = a \cos \Delta\omega t \sin \phi + b \sin \Delta\omega t \cos \phi \quad . \quad . \quad (14)$$

$$y = 1 + a \cos \Delta\omega t \cos \phi - b \sin \Delta\omega t \sin \phi \quad (15)$$

Appendix II shows that an approximate value for the distortion and demodulation caused by both magnitude and phase asymmetry of counterpart sidebands may be obtained by writing

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* *Wireless Engineer*, 1936, vol. 13, pp. 517, 534.

* *Wireless Engineer*, 1936, vol. 13, p. 534.

$$(x^2 + y^2)^{\frac{1}{2}} \simeq y\left(1 + \frac{x^2}{2y^2}\right) = y + \frac{x^2}{2y} \simeq y + \frac{x^2}{2}$$

There are thus two approximations: one is based upon the fact that $\sqrt{(1 + Z)} \simeq 1 + \frac{1}{2}Z$, where Z is a quantity not much greater than 1; the other is to assume that y in the denominator of the fraction $x^2/(2y)$ is always equal to the unmodulated carrier amplitude (assumed to be unity). It will be seen that, in the majority of cases, x does not in fact exceed y. Consideration also shows that the assumption that y can be written as unity in the ratio $x^2/(2y)$ is justified in the majority of cases because, where FC does in fact greatly exceed OF (Fig. 5) and where, therefore, y is considerably greater or less than unity, x is very small and therefore

$$y + \frac{x^2}{2y} \simeq y + \frac{x^2}{2}$$

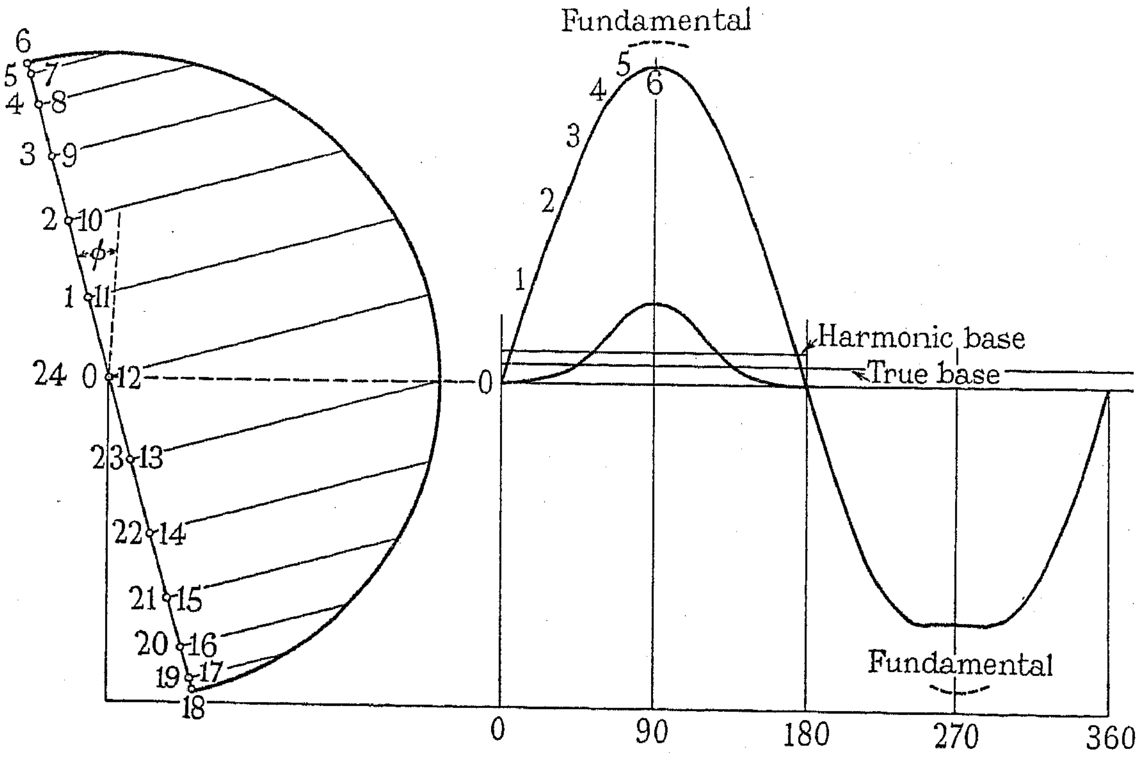

Fig. 7

$M_N = 1$, $\xi = 1$, $2\phi = 30°$, $M_{A\phi} = 0{\cdot}95$, $\delta_\phi = 7{\cdot}4$ %.

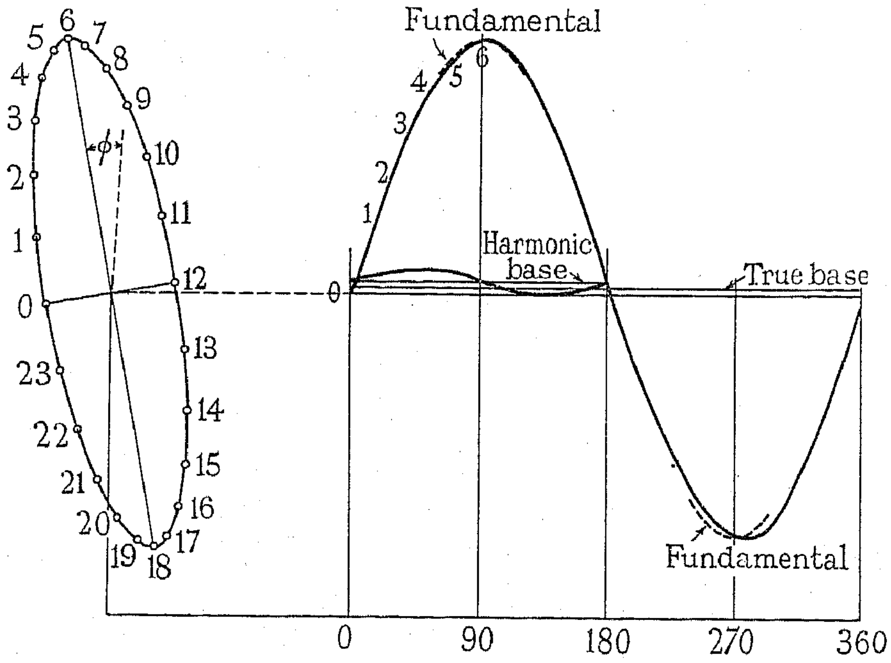

Fig. 8

$M_N = 1$, $\xi = 0{\cdot}6$, $2\phi = 20°$, $M_A = 0{\cdot}77$, $\delta = 2{\cdot}8$ %.

There are, however, possible conditions where the approximation cannot be justified, and these occur when ξ is unity or nearly unity, ϕ is of moderate value, and M_N is large. But if ξ and M_N have the value unity and ϕ is large, the approximation gives a true result just as it does if, whatever the values of ξ and M_N, ϕ is zero, i.e. the approximation is justified in the case where only magnitude asymmetry exists.

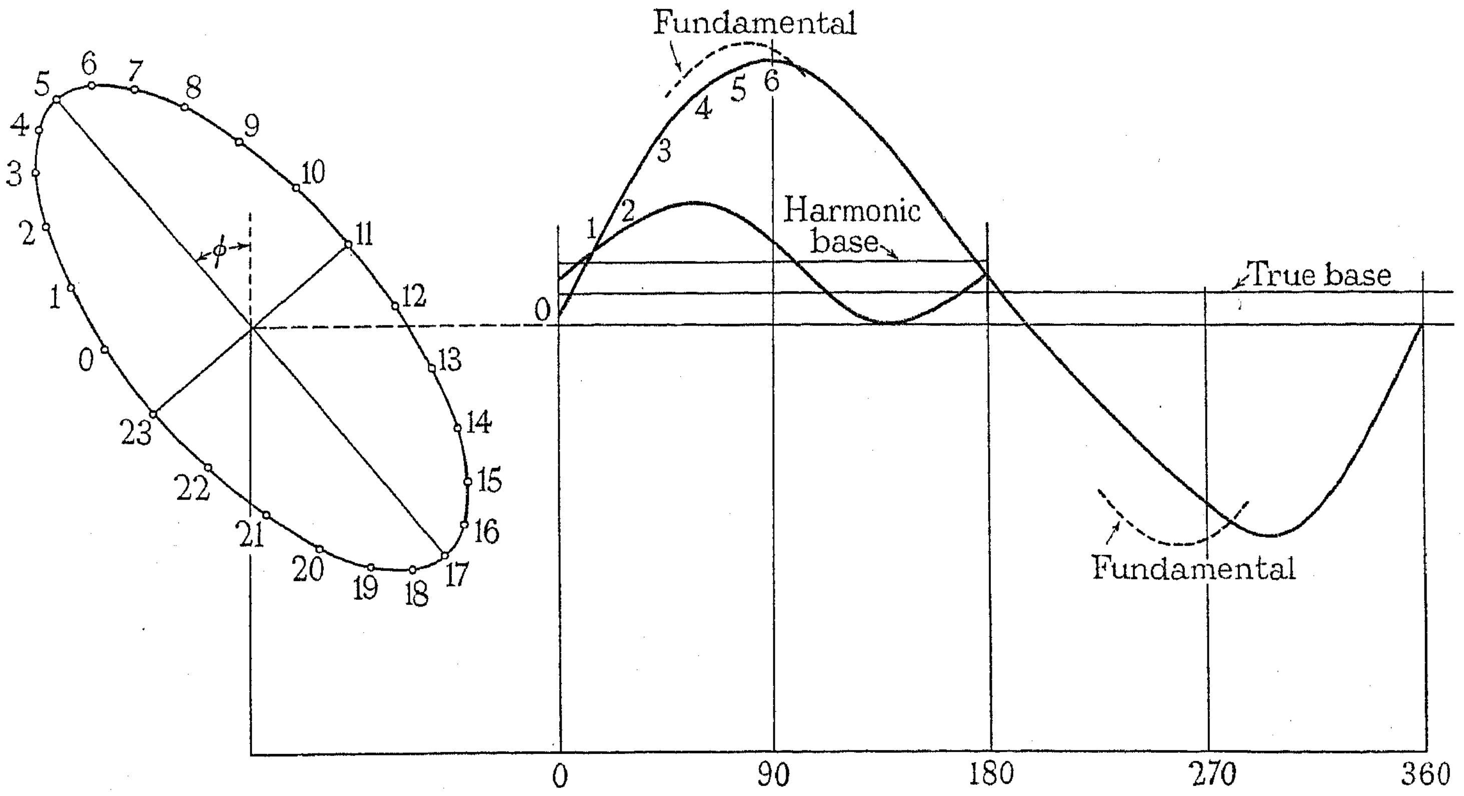

Fig. 9

$M_N = 1$, $\xi = 0\cdot4$, $2\phi = 80°$, $M_A = 0\cdot585$, $\delta = 12$ %.

The accuracy of the formula derived from the approximation can be judged from a study of Table 1, which is referred to again hereafter. It should, lastly, be noted that in the practice of asymmetric technique the approximation is practically always justified.

and that the modulation factor is given as

$$M_A = \frac{M_N}{2}(1 + \xi^2 + 2\xi \cos 2\phi)^{\frac{1}{2}} \quad . \quad . \quad (17)$$

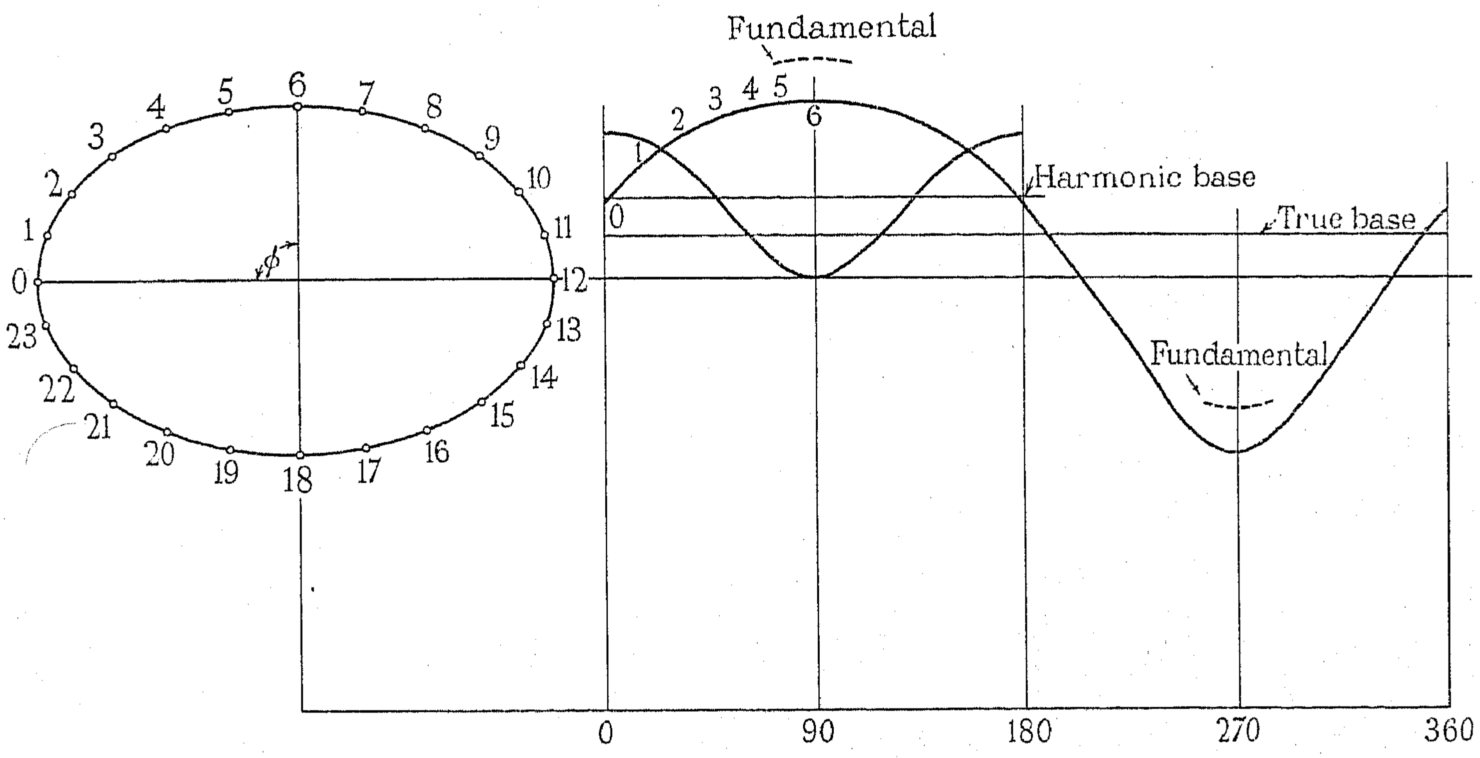

Fig. 10

$M_N = 1$, $\xi = 0\cdot2$, $2\phi = 180°$, $M_A = 0\cdot4$, $\delta = 24\cdot6$ %.

Thus Appendix II shows that the general case of asymmetric transmission can be analysed by writing that, if $\xi < 1$ and ϕ is finite, the distortion is given by

$$\delta = \frac{M_N}{8} \frac{1 + \xi^2 - 2\xi \cos 2\phi}{(1 + \xi^2 + 2\xi \cos 2\phi)^{\frac{1}{2}}} \quad . \quad . \quad (16)$$

The expressions for the special cases can be written as follows:—

When $\xi < 1$, $\phi = 0$ (magnitude asymmetry only),

$$\delta_\xi = \frac{M_N}{8} \frac{(1 - \xi)^2}{(1 + \xi)} \quad . \quad . \quad . \quad . \quad (18)$$

(see Appendix I)

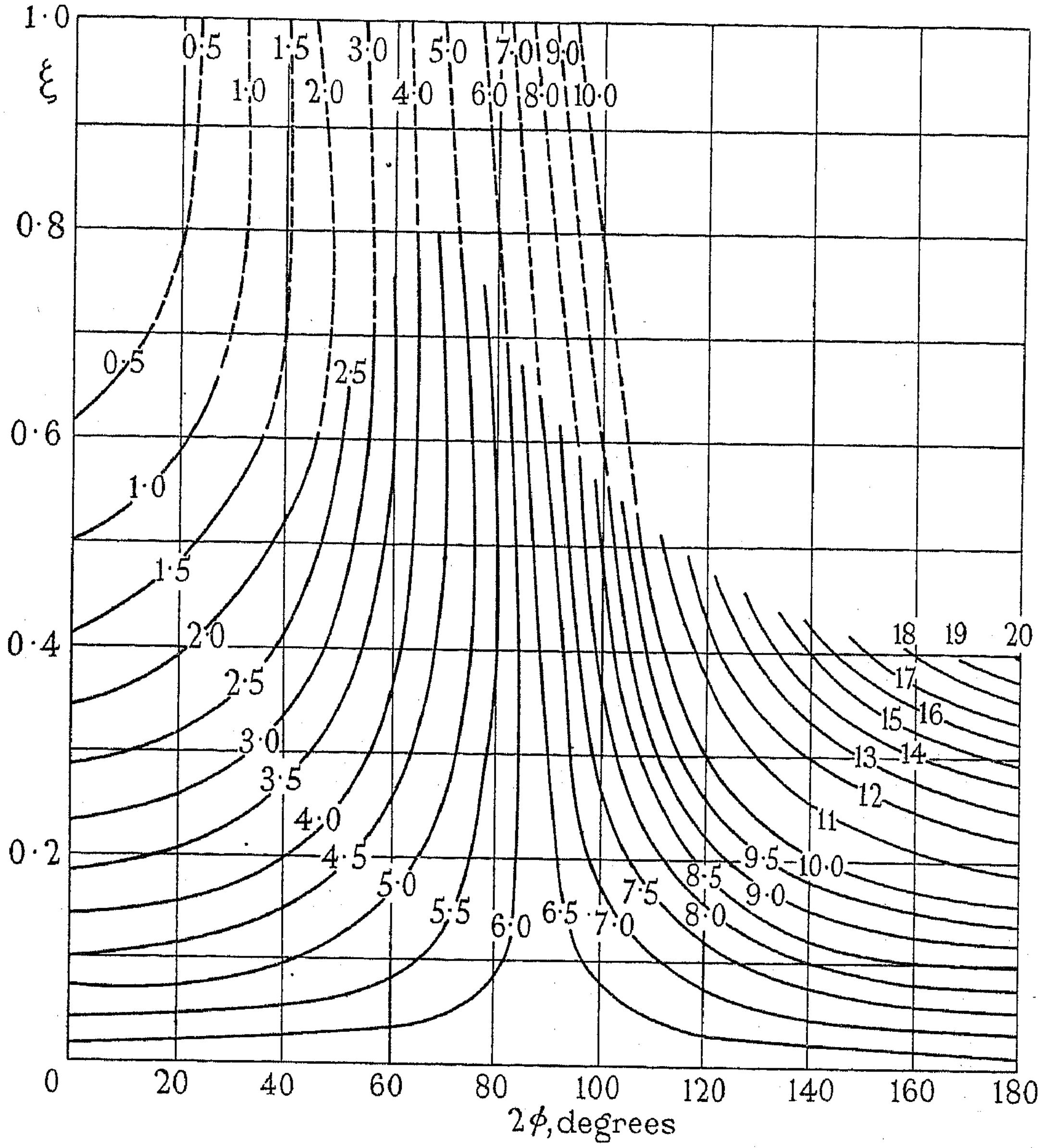

Fig. 11A

$M_N = 0{\cdot}5$. Figures on curves denote values of δ, expressed as percentages.

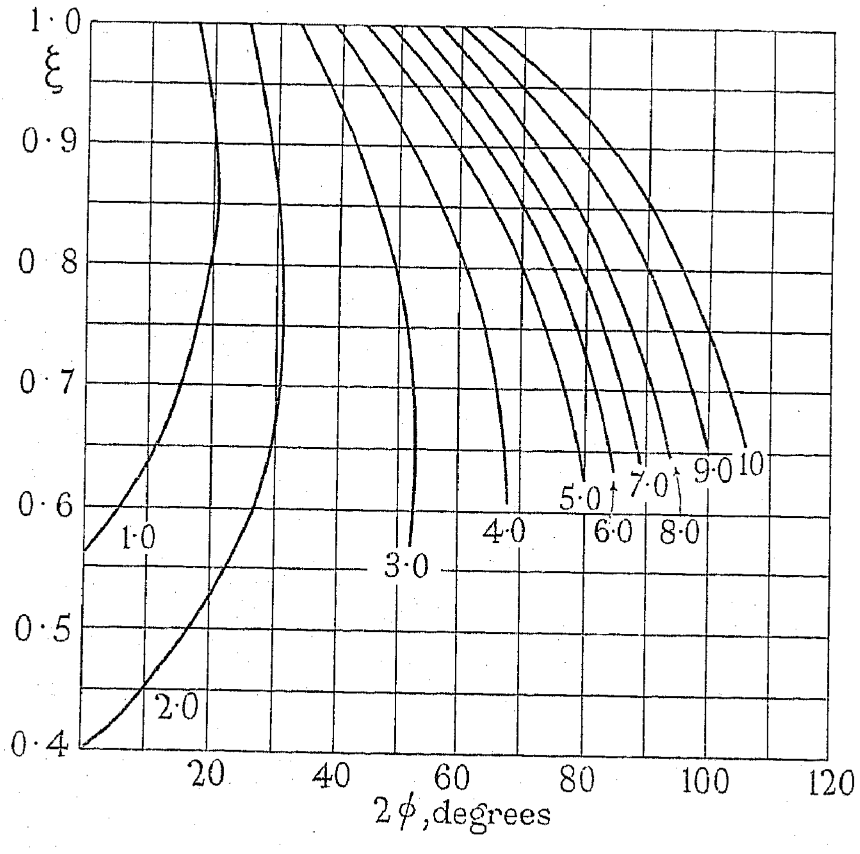

Fig. 11B

$M_N = 0{\cdot}75$. Figures on curves denote values of δ, expressed as percentages.

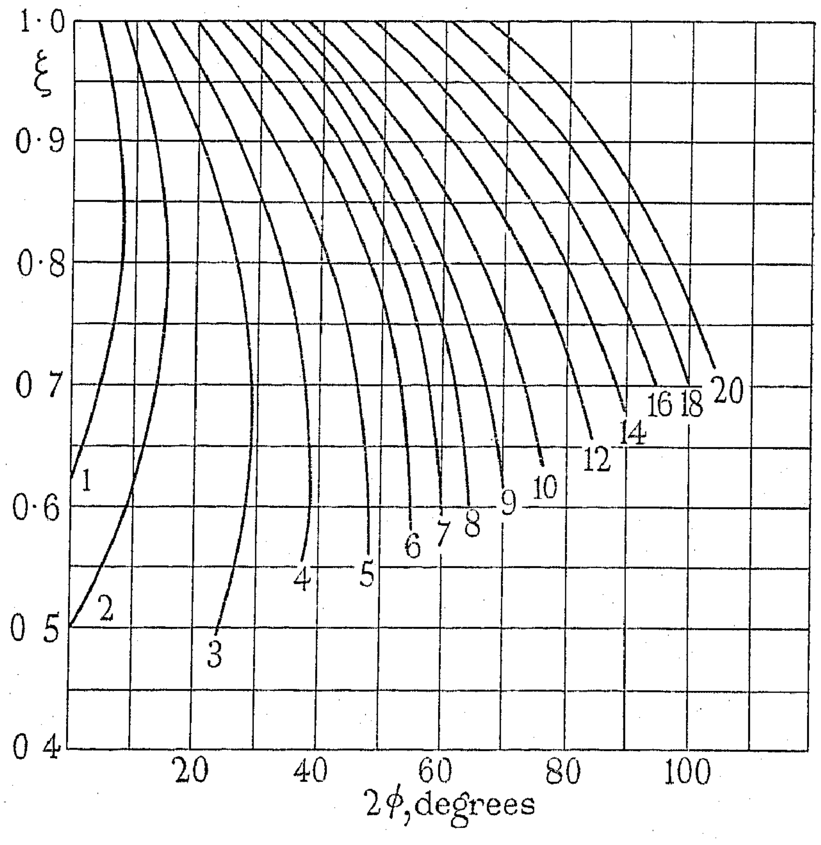

Fig. 11C

$M_N = 1$. Figures on curves denote values of δ, expressed as percentages.

$$M_{A\xi} = \frac{M_N}{2}(1+\xi) \quad . \quad . \quad . \quad . \quad . \quad (19)$$

[see equation (9)]

When $\xi = 1$, ϕ is finite (phase asymmetry only),

$$\delta_\phi = \frac{M_N\sqrt{2}}{8}\frac{(1-\cos 2\phi)}{(1+\cos 2\phi)^{\frac{1}{2}}} \quad . \quad . \quad . \quad (20)$$

$$M_{A\phi} = M_N\sqrt{2}(1+\cos 2\phi)^{\frac{1}{2}} \quad . \quad . \quad . \quad (21)$$

It is relevant to note in (21) that, if $\phi = 90°$ $(\theta_1 + \theta_2 = 180°)$, $M_{A\phi} = 0$; showing, what has been previously mentioned, that, in this condition, the carrier is completely demodulated. Since this fact can be proved from (12) and (13), which are not approximations, evidently (20) and (21) are accurate whatever the values of ξ and M_N provided that ϕ is large.

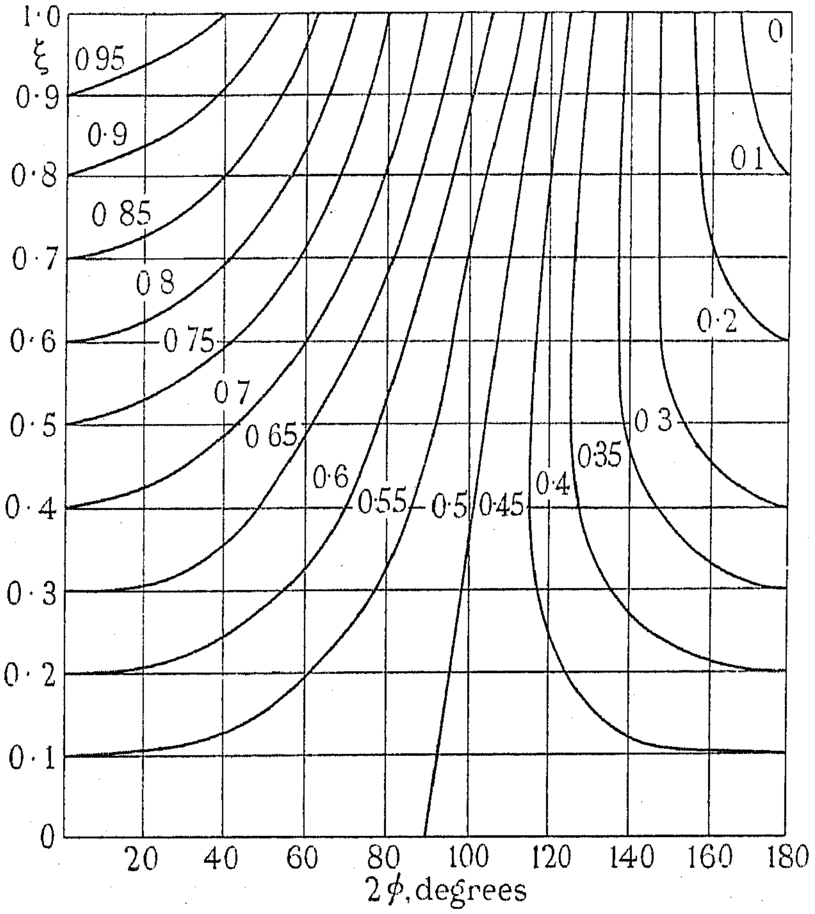

Fig. 12

$M_N = 1$. Figures on curves denote values of M_A.

In order, however, to supplement the information given by equations (16) to (21), derived from the approximations, it has been thought advisable to determine the degree of distortion when $\xi = 1$ and ϕ is finite and of medium value. It is a pity that the analysis given in the paper by Johnstone and Wright cannot be used, but, as has been pointed out,* it does not appear to check with straightforward graphical analysis. The author felt that in these circumstances it would be best to resort to a graphical method to derive the distortion.† The results are shown in Figs. 13A and 13B, and prove (Fig. 13A) that, when $\xi = 1$ and ϕ is finite, for values of modulation below 50 % the distortion is proportional to modulation, and (Fig. 13B) that the approximate equation (20) can be used for medium values of ϕ without introducing an error much above 10 % provided the modulation is less than 50 %.

A great many cases were analysed by graphical methods in the way illustrated in Figs. 6–10 inclusive, and Table 1 gives a comparison of the results (for $M_N = 1$) with those derived from equations (16) and (17). It is seen that the formula is accurate within the limits given.

Analysis of the Sound Spectrum

The conception of the asymmetric system is based upon the assumption that distortion may be kept constant and negligible, because, as asymmetry is increased between sideband components by the sideband-cutting filter, the modulation factor, on account of the nature of the sounds transmitted, becomes smaller. The foregoing expressions relating distortion and demodulation with phase and magnitude asymmetry would be useless in determining the requirements of the sideband-cutting filter unless supplemented by a knowledge of the relationship between M_N and the modulation frequency Δf.

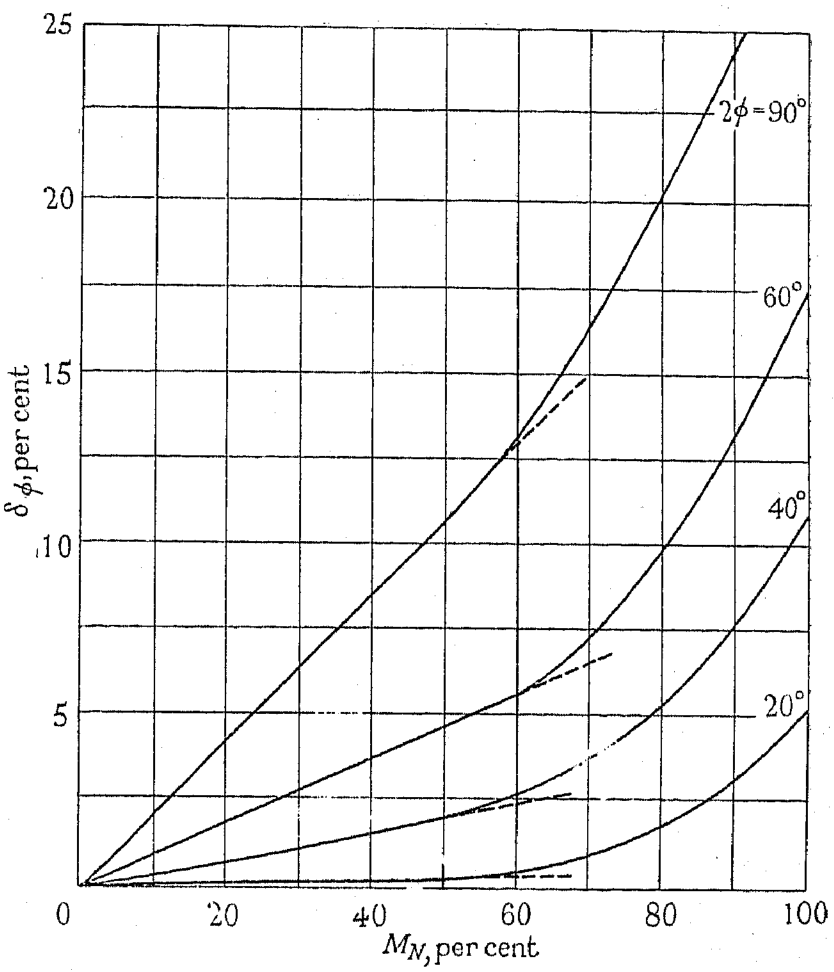

Fig. 13A

A great deal of work has been done in attempting to determine the nature of the sound spectrum. It is not proposed here either to discuss the methods adopted by other workers or necessarily to use the results so obtained. What we require to know in designing the asymmetric system is more a probability of the distribution of the magnitudes of the components within the spectrum of typical speech and music sounds than to be influenced by exceptional conditions.

The author some 7 years ago devised a method for analysing this probability. The circuits used comprised a resistance-coupled tuned circuit tuned to different audio frequencies, and the momentary peak maxima developed across this circuit, when energized from a microphone output, were measured by a "slide back" voltmeter. The measurements were repeated in 1936 and gave substantial agreement with the earlier results. Hence the curve of Fig. 14, derived from an average of hundreds of observations, is taken as a guide for designing asymmetric-sideband filters.

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* *Wireless Engineer*, 1936, vol. 13, p. 517.

† An approximate formula, given in Appendix I, applies when ϕ is of moderate values and M_N is large (see Fig. 13B).

It is not said that the curve is truly representative of the possible maxima which may on rare occasions occur, but it is argued that if it is used to guide the design of the system under discussion it is unlikely that such a system will give, over a long period of listening, any distortions which could be considered intolerable. Indeed, the choice of the values in Fig. 14 can only be criticized in terms of the performance of the asymmetric transmission system, rather than in terms of its comparison with other results obtained with a different object in view.

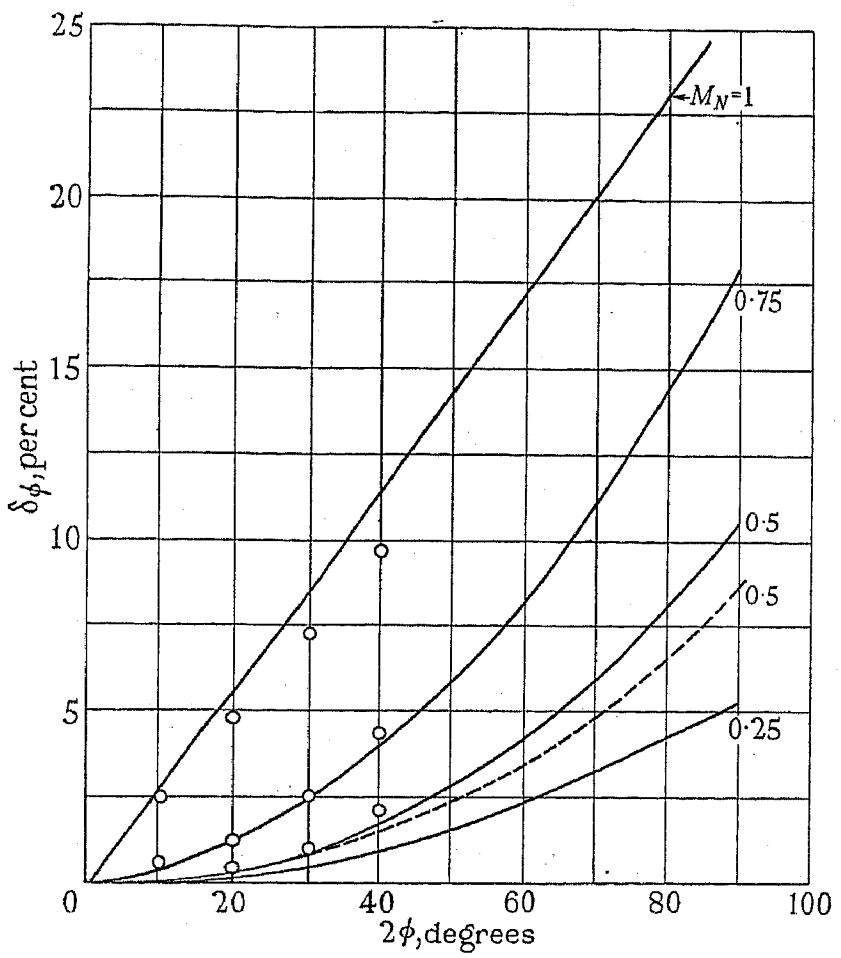

Fig. 13B

o o o Points derived from expression in Appendix II.
———— Graphical analysis.
- - - - - - From approximate formula (16).

Summary of the Analysis

The analysis shows that both phase and magnitude asymmetry between counterpart sideband components created by any given modulation frequency produce a distortion of the modulated-wave envelope and that this distortion takes chiefly the form of a 2nd harmonic of the fundamental.

The conception of the asymmetric systems implies that magnitude asymmetry between sideband components must be present. The analysis, however, reveals a point of major importance, namely that if both phase and magnitude asymmetry exist together in considerable degree, then the distortion will be very much greater than if magnitude asymmetry existed alone. It is practically impossible in the asymmetric system for phase asymmetry to exist without some magnitude asymmetry, but evidently phase asymmetry is dangerous when magnitude asymmetry is not very pronounced. Phase asymmetry may introduce a very serious demodulation of the carrier, a demodulation which could be, even in practical conditions of working, far greater than that introduced by magnitude asymmetry. It will be shown later, however, that it is possible to avoid the effects of phase asymmetry and yet produce a very large attenuation of a part of one sideband.

Curves relating ξ and ϕ for constant values of δ.

The summary to this analysis is given in a form convenient for direct consultation by engineers engaged in designing asymmetric-sideband systems. This summary is to be found in Figs. 11A, 11B, and 11C, where curves of constant distortion are plotted against ξ and 2ϕ.

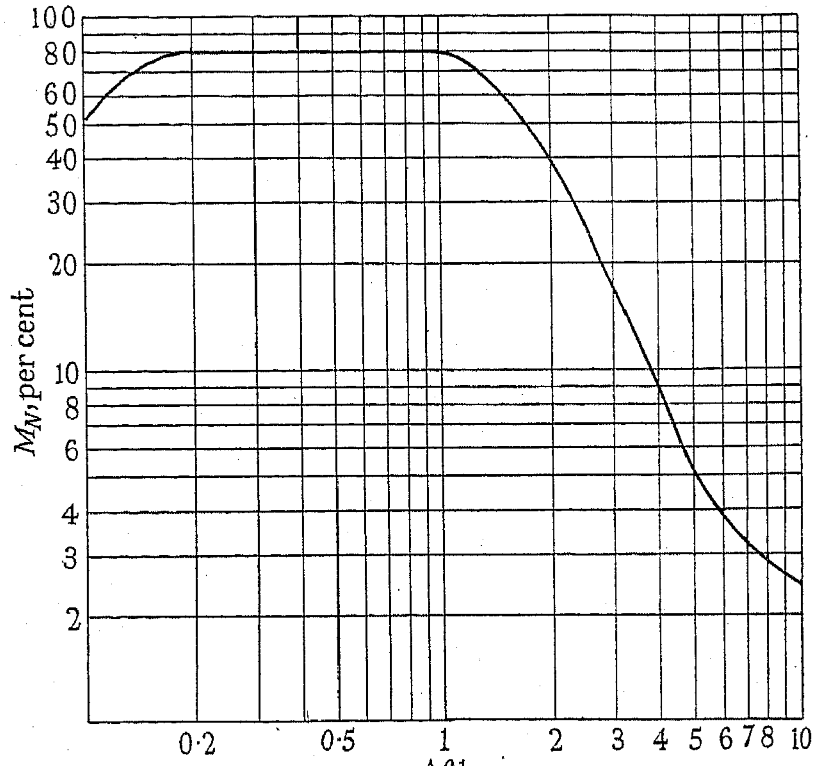

Fig. 14.—Analysis of sound spectrum (author's results).

Fig. 11A gives a set of curves derived from equation (16) for $M_N = 0 \cdot 5$ (50 % modulation). Values are only given which are likely to have practical relevance to asymmetric technique. The full-line curves represent the cases where the approximate formula applies for all values of M_N, and the broken-line curves apply for all values of M_N below and including $0 \cdot 5$. The distortion can thus be assessed for any value of M_N, if the full-line curves are used, by multiplying the distortion found from Fig. 11A by the new modulation factor, at which it is required to find the distortion, divided by $0 \cdot 5$. The dotted lines are only used if M_N exceeds $0 \cdot 5$. Figs. 11B and 11C show the curves which apply where

δ is not always proportional to M_N. The curves are drawn for modulation factors of 0·75 and 1·0 respectively. These curves are derived from the information given in Figs. 13A and 13B, which were in turn derived from a graphical analysis. The charts, taken together, give all the required information and meet all practical cases.

The points on the curves of Fig. 11A where $\xi < 1$ and ϕ is finite were found by plotting curves of δ against ϕ for different values of ξ. The values of ξ and ϕ for constant values of δ were then read off directly and transferred to Fig. 11A. This simple method was adopted because the conversion of (16) to express ξ in terms of the constants M_N and δ and the variable ϕ (or vice versa to express ϕ in terms of the constants M_N and δ and the variable ξ) involves rather heavy algebra and it was thought that the use of curves would be quicker, simpler, and probably more reliable.

Table 1

For $M_N = 1$

Values of ξ and 2ϕ		Results of graphical analysis		Results from chart of Figs. 11 and 12		Ratio of results of graphical analysis to those from charts	
ξ	2ϕ	δ	M_A	δ	M_A	δ	M_A
1	30	7·4	0·95	2*	0·97	3·7	0·98
0·8	20	2·8	0·88	0·8*	0·875	3·5	1·005
0·8	40	6·8	0·89	3·0*	0·85	2·26	1·05
0·6	20	2·8	0·77	1·8	0·78	1·55	0·99
0·6	40	4·0	0·77	5·5	0·75	0·73	1·025
0·6	100	14·2	0·6	18·5	0·52	0·77	1·15
0·4	40	6·3	0·66	5·0	0·66	1·26	1·00
0·4	80	12·0	0·585	11·1	0·57	1·08	1·02
0·4	100	16·0	0·52	16·0	0·51	1·00	1·02
0·2	0	7·0	0·6	6·8	0·6	1·03	1·00
0·2	40	7·3	0·565	8·0	0·57	0·91	0·99
0·2	90	12·1	0·495	13·2	0·51	0·92	0·97
0·2	180	24·6	0·4	22·3	0·4	1·1	1·00

* These values are for cases where the approximation is no longer justified and where Figs. 11B and 11C must be used. Note the good agreement for M_A, which does not depend upon an approximation.

Certain fixed points can, however, be obtained by giving ϕ certain specific values and then expressing ξ in terms of M_N and δ. Thus when $\phi = 0$ it can be shown from (18) that ξ is given by

$$(\xi)_{\phi=0} = (1 + d_1) - \left[d_1(d_1 + 4)\right]^{\frac{1}{2}} \quad . \quad . \quad (22)$$

where $\qquad d_1 = 4\delta_\xi/M_N$

and from (16) that, when $2\phi = 180°$,

$$(\xi)_{2\phi=180°} = \left[d_2(d_2 + 4)\right]^{\frac{1}{2}} - (1 + d_2) \quad . \quad (23)$$

where $d_2 = 4\delta_\phi/M_N$.

Points derived from (22) and (23) can be plotted directly on the charts without reference to curves. We may also convert (20) for Fig. 11A to give $\cos\phi$ in terms of M_N and δ. If $d_3 = 4\delta_\phi/M_N$, then

$$2\phi = \text{arc cos}\left[1 + d_3^2 - d_3(d_3^2 + 4)^{\frac{1}{2}}\right] \quad . \quad (24)$$

Curves relating ξ and ϕ for constant values of M_A.

The degree of demodulation of the carrier, given certain values of magnitude and phase asymmetry, can be readily found from the chart of Fig. 12. In order to plot the required curves we may resolve (17) and show that

$$\xi = -\cos 2\phi \pm \left[\cos^2 2\phi + (4M_A^2 - 1)\right]^{\frac{1}{2}}$$

and so give M_A constant chosen values and find ξ for a given value of 2ϕ. Evidently ξ must lie between 0 and 1. It can have alternative values of

$$\xi = -\cos 2\phi + \left[\cos^2 2\phi + (4M_A^2 - 1)\right]^{\frac{1}{2}}$$

and $\qquad \xi = -\cos 2\phi - \left[\cos^2 2\phi + (4M_A^2 - 1)\right]^{\frac{1}{2}}$

provided always the resulting values lie between 0 and 1. If $M_A = 0·5$, then

$$(\xi)_{M_A=0·5} = -2\cos 2\phi$$

so that, to be real, 2ϕ must exceed 90°. Thus when

$$2\phi = 90°, \ (\xi)_{M_A=0·5} = 0;$$

and when $\qquad 2\phi = 120°, \ (\xi)_{M_A=0·5} = 1.$

(4) DESIGN OF ASYMMETRIC-SIDEBAND TRANSMISSION SYSTEM

Asymmetric Carrier-Frequency Filter

The foregoing quantitative analysis furnishes all the information necessary to enable an expression to be determined for the required attenuation/frequency characteristic of a sideband-cutting filter which would give a constant and, it is believed, negligible distortion.

The increasing asymmetry factor as the modulation frequency increases is obviously a measure of increasing attenuation and distortion. But as the modulation frequency increases the modulation factor decreases, and so distortion is kept constant and negligible.

If ξ is known for any given modulation frequency (and hence sideband-component frequency), then, assuming that one sideband component is fully transmitted, α (the attenuation constant), which is a function of the reciprocal of ξ, will also be known. Since expressions exist to relate ξ with δ and M_N, therefore it is possible to find a value of α which, with a given value of M_N, keeps δ constant for any given sideband frequency.

It has been shown that, if phase asymmetry is neglected, then

$$\delta_{\xi} = \frac{M_N}{8} \frac{(1 - \xi)^2}{1 + \xi}$$

This equation must be converted so that ξ is given as a function of M_N and $\delta_{\xi 0}$; the latter being a constant, small, and negligible value of δ_{ξ}. This will give a value of ξ, at any particular modulation frequency, given by the value of M_N at that modulation frequency and the constant $\delta_{\xi 0}$. The values of ξ required at any given modulation frequency can then be converted to the values of α at the corresponding sideband frequency. Such values of α plotted against sideband frequency will then give the attenuation curve of a filter producing a constant distortion $\delta_{\xi 0}$.

The required curve having been found, the procedure will then be to choose a network giving as nearly as possible the required attenuation/frequency characteristic. The phase asymmetry introduced by such a network can be assessed either by calculation or by experiment. This will give values of both ξ and ϕ, and, from the charts of Figs. 11 and 12, the total distortion and demodulation can be found. If phase asymmetry appears to add more than a tolerable amount of distortion to that inevitably existing owing to magnitude asymmetry, alternative networks must be studied until a satisfactory compromise is reached; design requirements are revealed through trial and error.

Equation (22) expresses ξ in terms of M_N and δ. This, however, applies to a system in which the overall characteristic of the transmitter would not be level, since as ξ decreases the effective modulation of the carrier also decreases. This demodulation must be corrected in some part of the system. Correction must take the form either of multiplying the input modulations or of multiplying the output from the detector by a factor M_N/M_A. Equation (22) expresses the case where no correction is applied.

It is, in theory, immaterial whether correction takes place in the receiver (as an audio compensation in the post-detector circuits) or in the transmitter. In practice, however, correction of the asymmetric filter demodulation must take place in the transmitter, otherwise existing receivers would remark the changes when any transmitter was adapted to the asymmetric system. It is, however, pertinent to observe that correction for demodulation makes it necessary to include more of the cut-away sideband than would be necessary if demodulation correction were not applied. In order to illustrate this point two cases are considered throughout the paper, namely one in which correction is applied and the other where it is assumed to be absent.

If demodulation correction is used, then the equation for $\delta_{\xi 0}$ is given by multiplying the value of δ_{ξ} obtained from (22) by $M_N/M_{A\xi}$.

Thus
$$\delta_{\xi 0} = \frac{M_N}{8} \times \frac{M_N}{M_{A\xi}} \frac{(1 - \xi)^2}{1 + \xi}$$

Or, from (19),

$$\delta_{\xi 0} = \frac{M_N}{4}\left(\frac{1 - \xi}{1 + \xi}\right)^2$$

If again we write $d = 4\delta_{\xi 0}/M_N$, then it can be proved that

$$\xi_T = \frac{(\sqrt{d} - 1)^2}{d - 1} \quad . \quad . \quad . \quad . \quad (25)$$

showing the value of $\xi (= \xi_T)$ when demodulation correction is applied.

Equation (22) expresses the value of δ when the frequency characteristic is not level. Converting (22) to give ξ_R,

$$\xi_R = (1 + d) - \left[d(d + 4)\right]^{\frac{1}{2}} \quad . \quad . \quad . \quad (26)$$

We may write that the attenuation constant of the sideband components of reduced amplitude is given in decibels by

$$\alpha = 20 \log_{10} \frac{1}{\xi}\ .$$

Or, from (25) and (26),

$$\alpha_T = 20 \log_{10} \frac{d - 1}{(\sqrt{d} - 1)^2} \quad . \quad . \quad . \quad (27)$$

$$\alpha_R = 20 \log_{10} \frac{1}{(1 + d) - \left[d(d + 4)\right]^{\frac{1}{2}}} \quad . \quad (28)$$

Equation (27) shows that as d gets nearer and nearer to unity, α_T approaches infinity. Equation (28) shows that when $d = 0 \cdot 5$, α_R is infinity.

Thus it is seen that, with demodulation correction, distortion will not be increased above a certain value even though one side-band component is completely cut away, always provided that the modulation factor M_N does not exceed a value $4\delta_{\xi 0}$. Similarly, when the frequency characteristic is not level, the sideband may be wholly removed provided the modulation factor does not exceed $8\delta_{\xi 0}$. Writing the particular values of these modulation factors as $M_{N\infty T}$ and $M_{N\infty R}$ respectively, evidently $M_{N\infty R} = 2M_{N\infty T}$.

But M_N is a function of the modulating frequency Δf and decreases as Δf increases (see Fig. 14), so that $M_{N\infty T}$ may be related to a modulating frequency $\Delta f_{\infty T}$ and $M_{N\infty R}$ to a modulating frequency $\Delta f_{\infty R}$. Since $M_{N\infty R} = 2M_{N\infty T}$, therefore $\Delta f_{\infty R} < \Delta f_{\infty T}$. But the filter attenuation characteristic is a curve plotting α_T or α_R against carrier frequencies $(f_0 \pm \Delta f)$, and the frequencies above or below which the attenuation constants may be infinite may be expressed as $(f_0 \pm \Delta f_{\infty T})$ or $(f_0 \pm \Delta f_{\infty R})$, showing that since $\Delta f_{\infty R} < \Delta f_{\infty T}$ the frequency at and above or below which the attenuation may be infinite is closer to the carrier frequency f_0 if no demodulation correction takes place than when it is applied; or, by implication, that more of one sideband may be cut away if the frequency characteristic of the transmitter is allowed to remain unlevel than if it is arranged to be corrected.

This fact is obvious when it is realized that, to correct the frequency characteristic, M_N must be multiplied by $2/(1 + \xi)$, and that this increasing modulation of the carrier as ξ decreases must introduce a greater distortion which can only be compensated for by allowing for less asymmetry at a given modulation frequency than if M_N did not have to be increased above its normal value.

The curves in Figs. 15A and 15B show the values of attenuation plotted against sideband frequency (expressed as $\Delta f = f_0 - f_s$) for the two cases.

It will be appreciated that, if $\delta_{\xi 0}$ (and hence d) has a value greater than zero, then even at sideband-component frequencies infinitely close to the carrier frequency α also will be greater than zero. But since one set of sideband components is transmitted with zero attenuation the curve plotting the attenuation constant against frequency for the requirements implicit in (27) and (28) will show a discontinuity at the carrier frequency. This is because it has been assumed that some distortion is tolerable at all sideband frequencies. Actually, equations (27) and (28) should be considered to apply only for those sideband-component frequencies created by modulation components having frequencies greater than that (say 1 200 cycles per sec.) where modulation decreases with increasing frequency of modulation. In drawing the criterion attenuation/frequency characteristics for asymmetric-sideband carrier-frequency filters it will be necessary, therefore, only to use (27) and (28) for sideband-frequency values above or below ($f_0 \pm 1\,200$) cycles per sec., say, and to join the point showing the value of α_T or α_R for a frequency of ($f_0 \pm 1\,200$) cycles per sec. to a point representing f_0 and zero attenuation by a curve estimated to represent the probable curve obtained in practice. Curves derived from (27) and (28) and smoothed to fit practical working conditions are given in Figs. 15A and 15B. The broken lines are values given by (27) and (28) at the lower frequencies. It has been assumed that it is the lower-frequency sideband which is cut away; it might equally have been the higher.

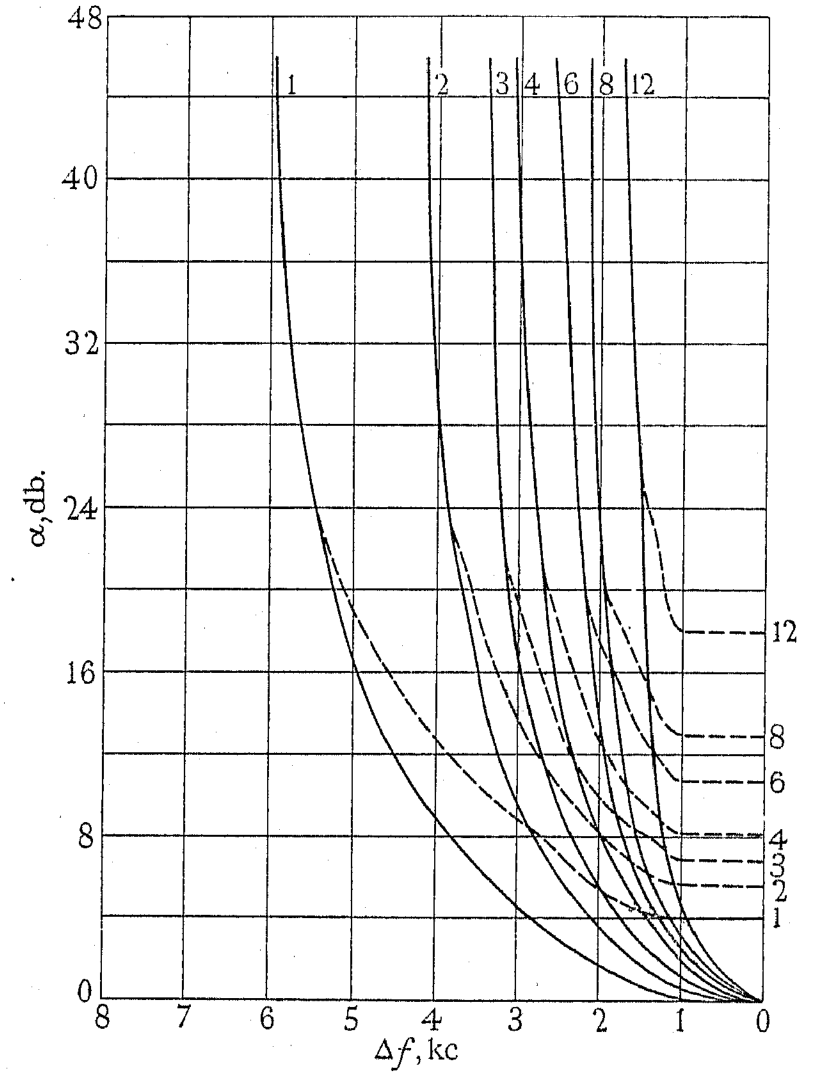

Fig. 15A.—Modulation-characteristic corrected.

Figures on curves denote values of δ_{ξ_0}, expressed as percentages.

——— Desired attenuation. - - - - - - From $\dfrac{1-d}{(\sqrt{d}-1)^2}$

The required attenuation curve of the band filter wherein the lower-frequency sideband is cut away is thus one in which attenuation should increase gradually from the cut-off frequency as the frequency decreases, but wherein, as this frequency decreases still further, the attenuation may increase very rapidly and below a certain side-band frequency may become infinite.

The problem of designing such a filter would be most nearly solved, so far as the production of the correct frequency/attenuation characteristic is concerned, by the use of a single 3- or 4-element section matched with a section having f_∞ a little lower than ($f_0 - \Delta f_\infty$). This solution fails, however, because the phase characteristic of such a combination would give intolerable phase-asymmetry distortion. A single section would tend to give the correct attenuation/frequency characteristic over a range of frequencies close to the carrier frequency but would not greatly attenuate the outer parts of one sideband, thus not being as effective as desirable in bringing out the possible merits of the system.

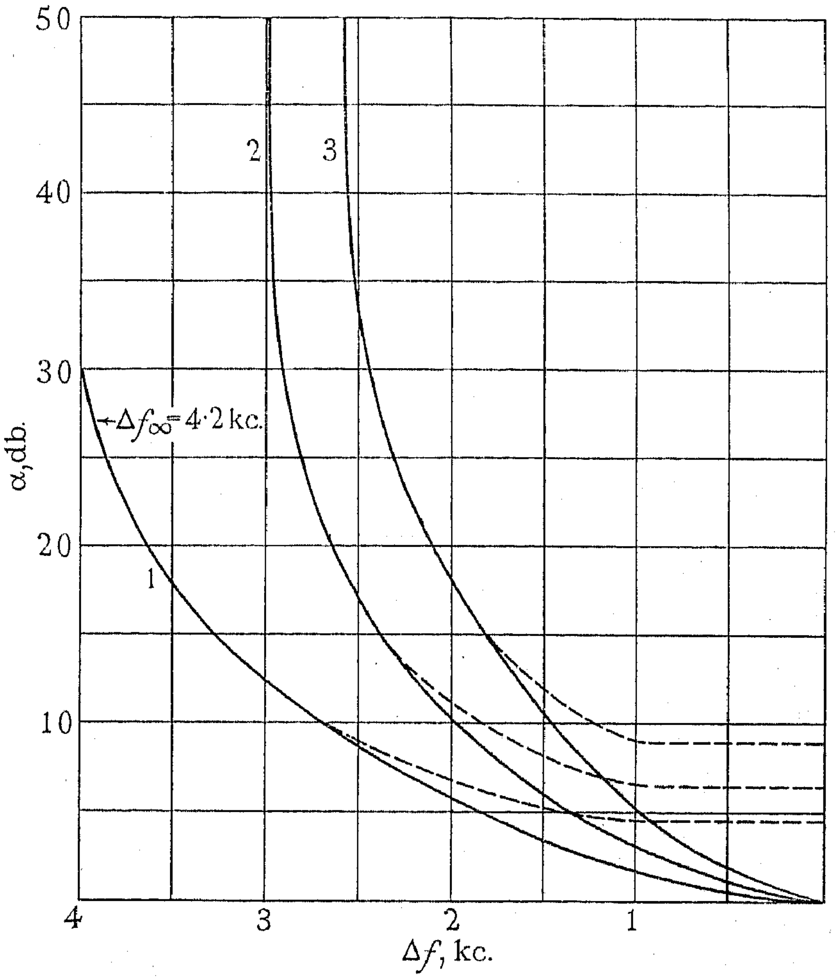

Fig. 15B.—Modulation-characteristic uncorrected.

Figures on curves denote values of δ_{ξ_0}, expressed as percentages.

——— Desired attenuation. - - - - - - From $\dfrac{1}{1+d} - [d(d+4)]^{\frac{1}{2}}$

It might be possible to produce the required characteristic by associating filter sections having "staggered" cut-off frequencies. According to this idea, and if the lower-frequency sideband were being attenuated, one filter section might have a lower cut-off frequency at f_0 kc./sec., the next at $(f_0 - 1)$ kc./sec., the next at $(f_0 - 2)$ kc./sec., and so on. It would be necessary, since their cut-off frequencies would be different, to decouple each section from the other, and this would be very simply done either by using decoupling-resistance networks matching or tending to match the filter-section image impedances, or by valves. Some amplification would in any case be necessary to compensate for the losses in the resistance networks, and therefore some sections could be coupled by valves directly and others by resistance networks. The phase characteristics of such combinations would tend to be more uniform than if an f_∞ section were used, because the carrier would be located nearer and nearer (in the case considered where the lower-frequency sideband was attenuated), in each successive filter, to f_m. A quantitative study of such a system shows, however, that while the expected tendencies are revealed the actual phase distortion becomes too large to be acceptable.

The preferred solution to the problems lies in using a constant-k band-pass filter section in combination with a constant-k band-elimination filter section. A network of this form is shown in Fig. 16. The resistance network, which is, more suitably, replaced by valves, is introduced in order that the two sections, having different image impedance/frequency characteristics, may be dissociated electrically from one another. The importance of the scheme lies in its facility to eliminate phase distortion and at the same time to give the correct form of frequency/attenuation characteristic.

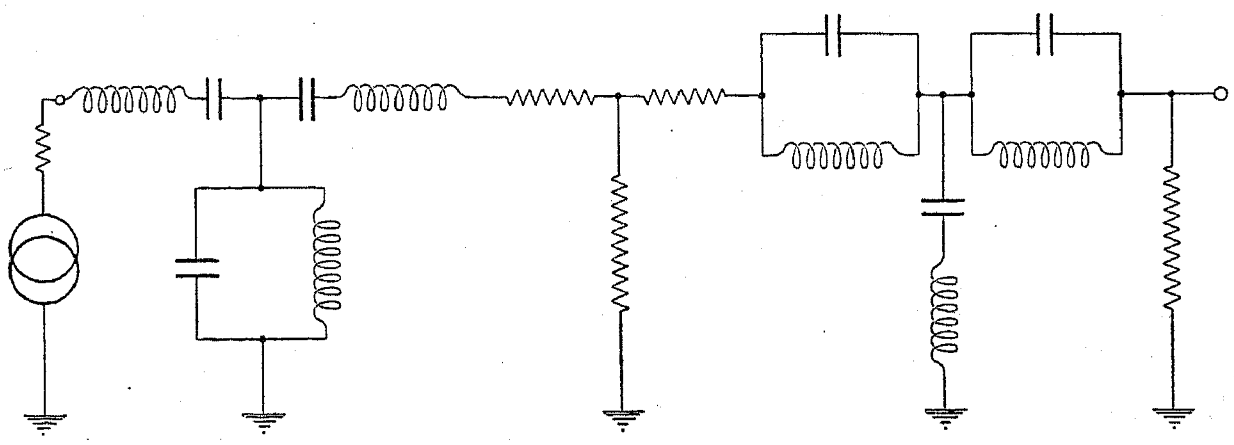

Fig. 16

Let θ be the phase angle between currents entering and leaving a filter. Let θ_0 be the value of θ at a carrier frequency f_0, and let θ_1 and θ_2 respectively be the values of θ for frequencies corresponding to any two sideband frequencies $(f_0 + \Delta fx)$ and $(f_0 - \Delta fx)$.

The requirement for phase symmetry between counterpart sideband components is that the total effect of the network shall be to make $(\theta_0 - \theta_1) = (\theta_2 - \theta_0)$, it being assumed that $\theta_2 > \theta_0 > \theta_1$. This condition is fulfilled if, in two networks, such as those shown in Fig. 16 and connected so that the final output is determined by their combined performance, the sum of the differentials expressing the rate of change of θ with frequency, at a number of frequencies in the working sideband range, is equal to a constant. This implies that either of the filters alone may have a flexure in its phase characteristic provided always the flexures in each filter are equal and opposite to one another.

Considering the case in which the lower sideband is attenuated, consultation of textbooks on wave filters will show that a single-section constant-k band-pass filter terminated in a resistance has a phase characteristic which, over the range of frequencies between its lower cut-off frequency f_{1b} and frequencies less than f_{1b}, has a flexure such that the rate of change of θ with frequency decreases as the frequency decreases. The curve thus flattens out as the frequency decreases, becoming asymptotic to a zero value of θ at zero frequency. In a constant-k band eliminator, however, the rate of change of θ with frequency increases as frequency is decreased from a value greater than $f_{2\epsilon}$ (the upper cut-off frequency of the band eliminator) to $f_{2\epsilon}$. Suppose the networks of Fig. 16 are arranged so that the carrier frequency f_0 is coincident with the lower cut-off frequency of the band filter and that the value $f_{m\epsilon}$ for the band eliminator approximates to $(f_0 - \Delta f_\infty)$ while the upper cut-off frequency $f_2{}^{\epsilon}$ of the band eliminator is a little greater than $f_{m\epsilon}$ but less than $(f_0 - \Delta f_\infty)$. Consideration will show that, given these conditions over a working range of frequencies f_0 to $(f_0 - f_{2\epsilon})$ and because the flexures of the two phase characteristics are inverse, the sum of their differential constants will tend to be a constant, i.e. the combination of the two characteristics tends to be a straight line of the form $y = kx$.

At frequencies less than $(f_0 - f_{2\epsilon})$, i.e. within the attenuation band of the band eliminator, very violent phase-changes take place and phase asymmetry is extremely pronounced. But obviously, in this band of frequencies, attenuation of one sideband component is so great, band eliminator and band pass combining their attenuations, that ξ is virtually zero and no added distortion due to phase asymmetry can take place. At frequencies greater than f_0 the rate of change of the band-eliminator phase characteristic tends to become very small while that of the band-pass filter is constant. Thus phase distortion is virtually eliminated by the combination of these inverse networks, while, thanks to the contributions made by the band eliminator to the attenuation given by the band-pass filter, the outer parts of one sideband are very greatly attenuated.

This result can be seen to be only obtainable by combining the performance of different types of filters having phase characteristics which are arranged to have inverse flexures over bands of frequencies wherein, in asymmetric technique, one sideband component is of magnitude comparable with its counterpart.

Equalizer Design

It has been shown that when a modulated carrier is passed through the sideband-cutting filter it suffers a demodulation at the output terminals of that filter. The demodulation ratio is given by M_A/M_N, or, if $M_N = 1$, as in Fig. 12, by M_A. Evidently the compensation required is to arrange for the input modulations to be multiplied by $1/M_A$. Let this multiplication factor be written $p = 1/M_A$. Now M_A has boundary values corresponding to $\xi = 1$ and $\xi = 0$ (low modulation frequency and high modulation frequency, respectively) of 1 and 0·5. Evidently, since M_A has values less than unity, p has values greater than unity, implying that the compensation circuits must embody an amplifier and, because M_A varies with modulation frequency, a passive frequency-discriminative network.

We may express this by writing $p = 1/(2M_A) \times m$, where m is the magnification constant of the required amplifier. If M_A is never less than 0·5 and m is constant and equal to 2, then $p \equiv 1/M_A$, which is the required condition. If M_A is, at any modulation frequency, less than 0·5, then, to maintain $p = 1/M_A$, m must exceed 2. But at the two boundary conditions $\xi = 1$, $\xi = 0$, it is seen that m must be equal to 2. This proves that if M_A ever falls below 0·5, a condition which can only arise if phase and magnitude asymmetry in the sideband components combine in certain proportions, the equalizer amplifier must be frequency-discriminative.

Thus the requirement of the equalizer circuit is that a passive network in association with an amplifier is required to be connected between the source of the audio-frequency modulations and the input to the modulating valve, or between the detector output and the loudspeaker, the network being characterized by an attenuation constant given by

$$\alpha_\epsilon = 20 \log_{10} (2M_A') \quad . \quad . \quad . \quad . \quad (29)$$

and an amplifier having a magnification constant given by $m = 1/M_A''$; where M_A' is any value of M_A exceeding or equal to 0·5 and being written 0·5 if M_A is less than 0·5, and M_A'' is any value of M_A less than or equal to 0·5 and being written 0·5 if M_A is greater than 0·5. The values of M_A can be found, given the values of ξ and ϕ, for any filter from the charts of Fig. 12. If phase asymmetry can be neglected,

$$\alpha_{\epsilon\xi} = 20 \log_{10} (1 + \xi)$$

and $m = 2$ always. This is the practical case because the filter network suitable for asymmetric technique should be designed to give, so far as possible, no phase asymmetry when ξ is large.

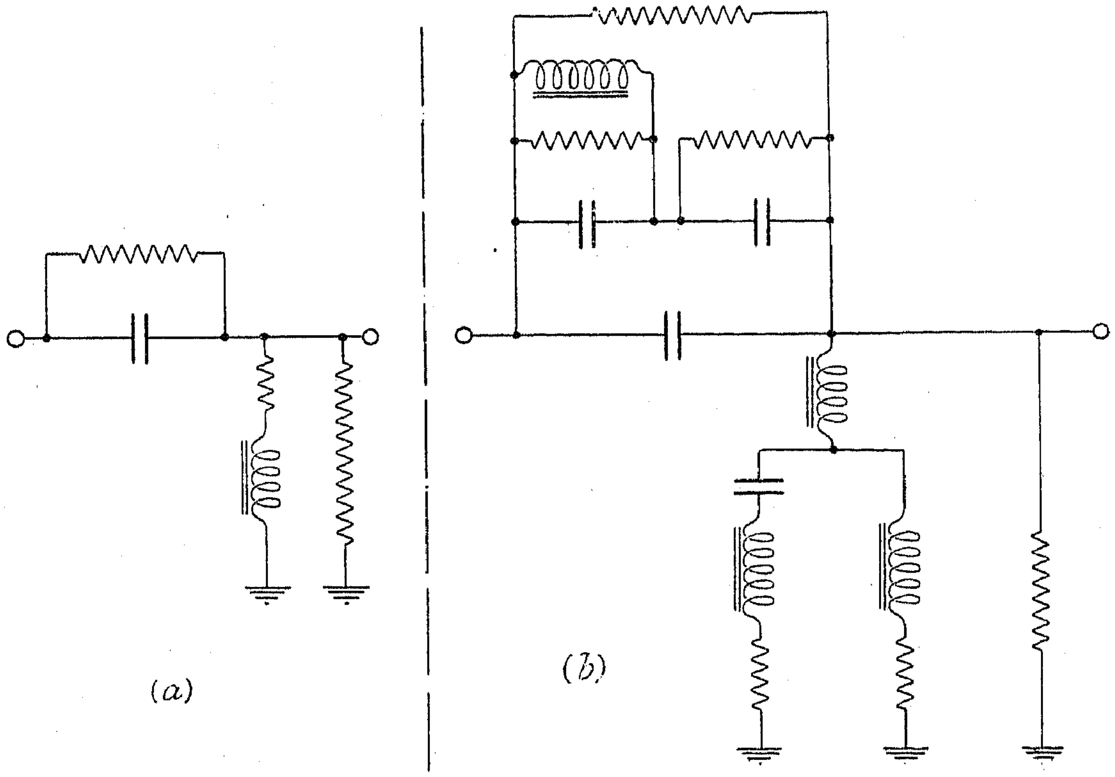

Fig. 17

A discussion of how networks, usually known as equalizers, are built up from simple elements is given in standard textbooks. It is essential that such networks should be designed for constant impedance, since they will be fed in practice from sources having by no means negligible internal impedance.

A simple form of network suitable in cases where demodulation of the carrier by the asymmetric filter is not excessive is shown in Fig. 17(*a*), while a more complex form used in certain experiments described later, where phase asymmetry was very pronounced, is shown in Fig. 17(*b*).

Practical Designs

It is proposed now to study different types of filter networks as regards their suitability for asymmetric-sideband transmission.

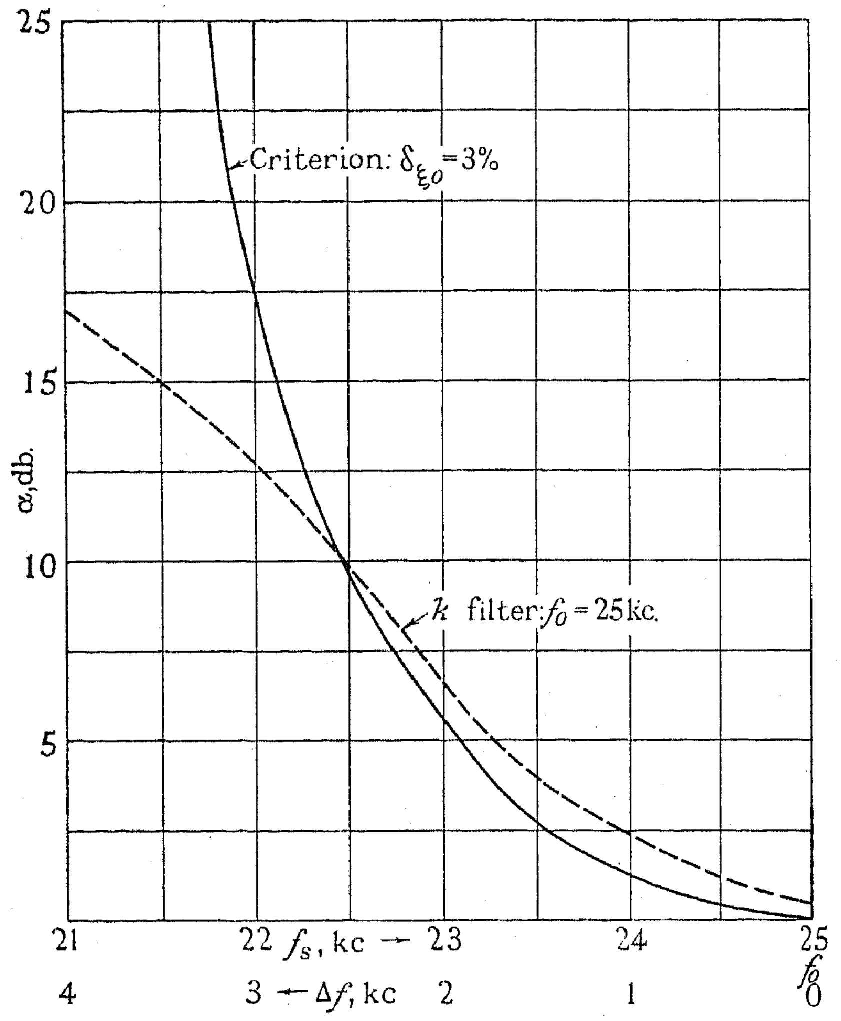

Fig. 18.—Constant-k single-section filter. Points obtained experimentally.

Constant-k single-section filter (f_0 = 25 kc.).

Fig. 1 shows the attenuation characteristic of a constant-k band-pass filter. Fig. 18 repeats Fig. 1 except that f_0 is chosen to have a frequency of 25 kc. instead of 27 kc. Fig. 18 compares the criterion attenuation curve for 3 % distortion (a figure chosen to represent maximum tolerable distortion) with the attenuation curve of the constant-k filter with the carrier located at 25 kc. Fig. 19 is derived from Fig. 1 and Fig. 18, and shows in full lines the values of ξ and ϕ for the conditions chosen for the particular filter under study.

We may now make a Table and derive the distortion and demodulation. Thus Table 2 shows in col. 1 the values of Δf and in col. 2 the corresponding values of M_N. This is obtained from Fig. 14. Cols. 3 and 4 respectively give the values of ξ (from Fig. 1 or Fig. 18) and ϕ (from Fig. 1), corresponding to the given values of Δf in col. 1.

Given ξ and ϕ, we can write in col. 5 the corresponding value of M_A using the chart of Fig. 12. Col. 6 gives M_{Nx}, the actual maximum value of modulation taking place, and is obtained by dividing M_{Np} (the maximum value of M_N input to the transmitter modulator terminals) by M_A. (This increase of M_N is brought about by the action of the equalizer compensating for the demodulation.) Cols. 3, 4, and 7 now give all the information necessary to obtain δ from the charts of Figs. 11A, 11B, or 11C. If ξ, ϕ, and M_{Nx} are of suitable values, Fig. 11A can be used directly; if not, Fig. 11B or Fig. 11C must be used. Table 2 is arranged to show which charts are used and how the distortion is derived. Fig. 19 is now

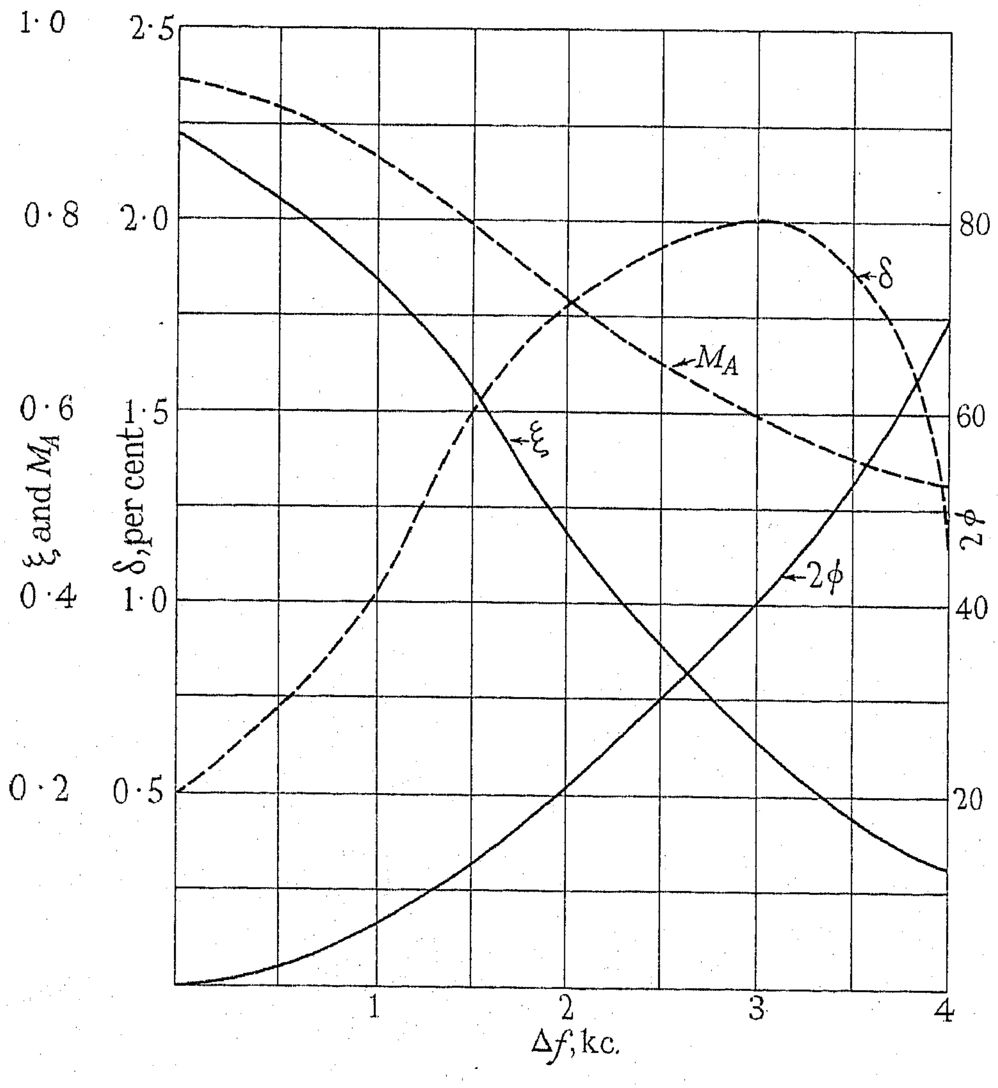

Fig. 19.—Constant-k single-section filter. f_0 = 25 kc.

Table 2

CONSTANT-k SECTION HAVING CHARACTERISTICS AS SHOWN IN FIG. 1 ($f_0 = 25$ KC./SEC.)

Δf (cycles/sec.)	M_{Np} (Fig. 14)	ξ (Fig. 1 or Fig. 18)	2ϕ (degrees) (Fig. 1)	M_A (Fig. 12)	M_{Nx}	δ		
						Chart	Chart reading	Actual value
1	2	3	4	5	6	7		
0	0·8	0·89	0	0·945	0·845	Fig. 11A	0·25	0·5
1 000	0·8	0·744	6	0·87	0·92	Fig. 11C	1·0	1·0
2 000	0·4	0·47	21	0·72	0·555	Fig. 11A	1·6	1·775
3 000	0·17	0·26	41	0·605	0·278	Fig. 11A	3·6	2·0
4 000	0·07	0·125	69	0·525	0·101	Fig. 11A	5·5	1·1
5 000	0·05	0·1	97	0·495	0·101	Fig. 11A	6·5	1·31

completed by drawing in the values of δ and M_A (in broken lines). The Figure now gives all the required information.

It is seen that the distortion, using the constant-k section, would never exceed 2 %, and that the demodulation effects are not so great as to make the equalizer complex. The sole disadvantage, but that a considerable one, is that the attenuation characteristic is far from ideal.

Constant-*k* section terminated by f_∞ section.

One method of improving the filter attenuation characteristic is to associate an f_∞ section with a constant-k type of filter; but this introduces phase asymmetry. The network performance will, however, be examined not only because it affords an interesting example to illustrate the theory but also because the filter examined was actually used in certain experiments described hereafter. The attenuation/frequency and phase/frequency characteristics are shown in Fig. 25. Fig. 21 gives in full lines the values of ξ and ϕ. Calculations exactly similar to those already described give M_A and δ.

It is seen from Fig. 20, which compares the criterion attenuation curve with that obtainable from the constant-k section terminated by the f_∞ section, that the attenuation characteristic given by the network approaches the ideal far more nearly than if (see Fig. 18) the k section is used alone. But it is also seen from Fig. 21 that the network, in achieving a better attenuation characteristic, has introduced intolerable distortions. These are remarkably concentrated between the modulation frequencies of 2 500 and 3 500 cycles per sec.

Constant-*k* band-pass section associated with constant-*k* band-elimination filter.

Fig. 22 gives the attenuation/frequency and phase/frequency characteristics of the network therein illustrated, namely the combination of two constant-k sections, one for band pass and the other for band elimination. The method employed to decouple the two networks is not shown because there are many alternative forms. The points are, however, taken experimentally.

It is at once seen that the phase characteristic is remarkably straight and the attenuation characteristic markedly steep. Fig. 23 compares the attenuation characteristic with the criterion. It will be shown later that the fact that the attenuation characteristic returns to only moderate values of attenuation after $f_s = 19$ kc./sec., or for a modulation frequency of $25 - 19 = 6$ kc./sec., does not greatly affect the efficacy of the network for preventing spectrum overlap. An f_∞ band-eliminator section could, however, be used.

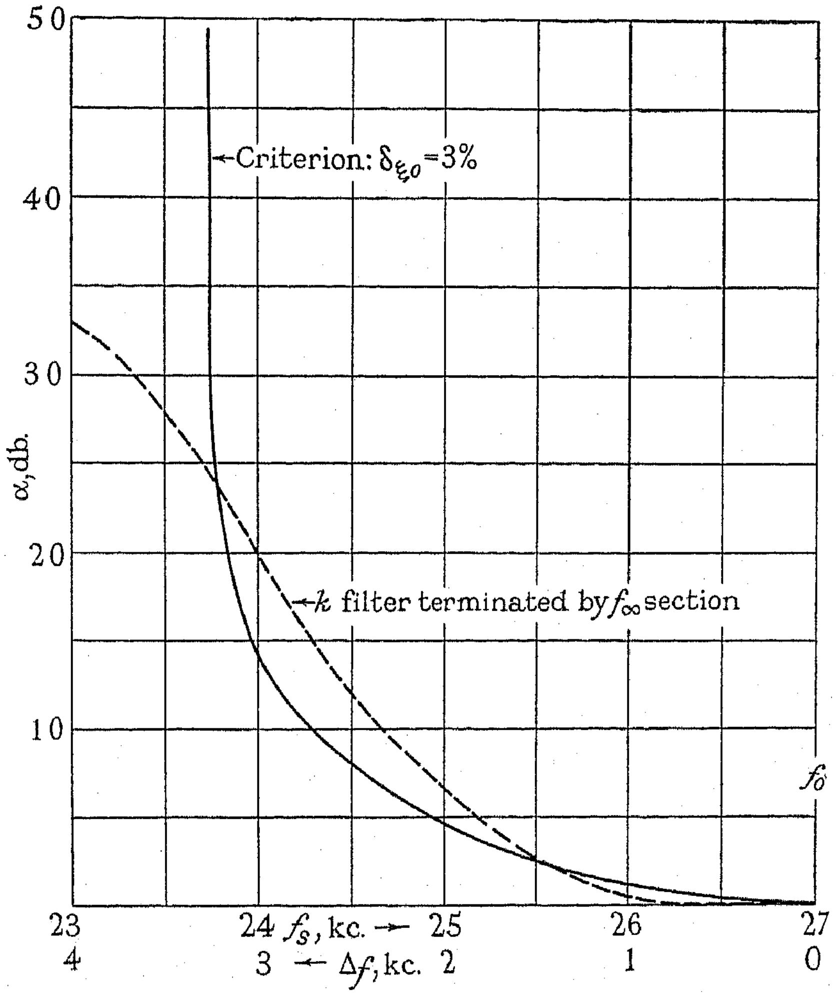

Fig. 20.—Filter comprising constant-k section terminated by f_∞ section. Points obtained experimentally. $f_0 = 27$ kc./sec., $f_\infty = 21$ kc./sec.

Fig. 24 shows ξ and ϕ and the fundamental quantities by which performance may be judged, namely M_A and δ. It is seen that the distortion never exceeds the chosen value, that demodulation never exceeds 6 db., and yet

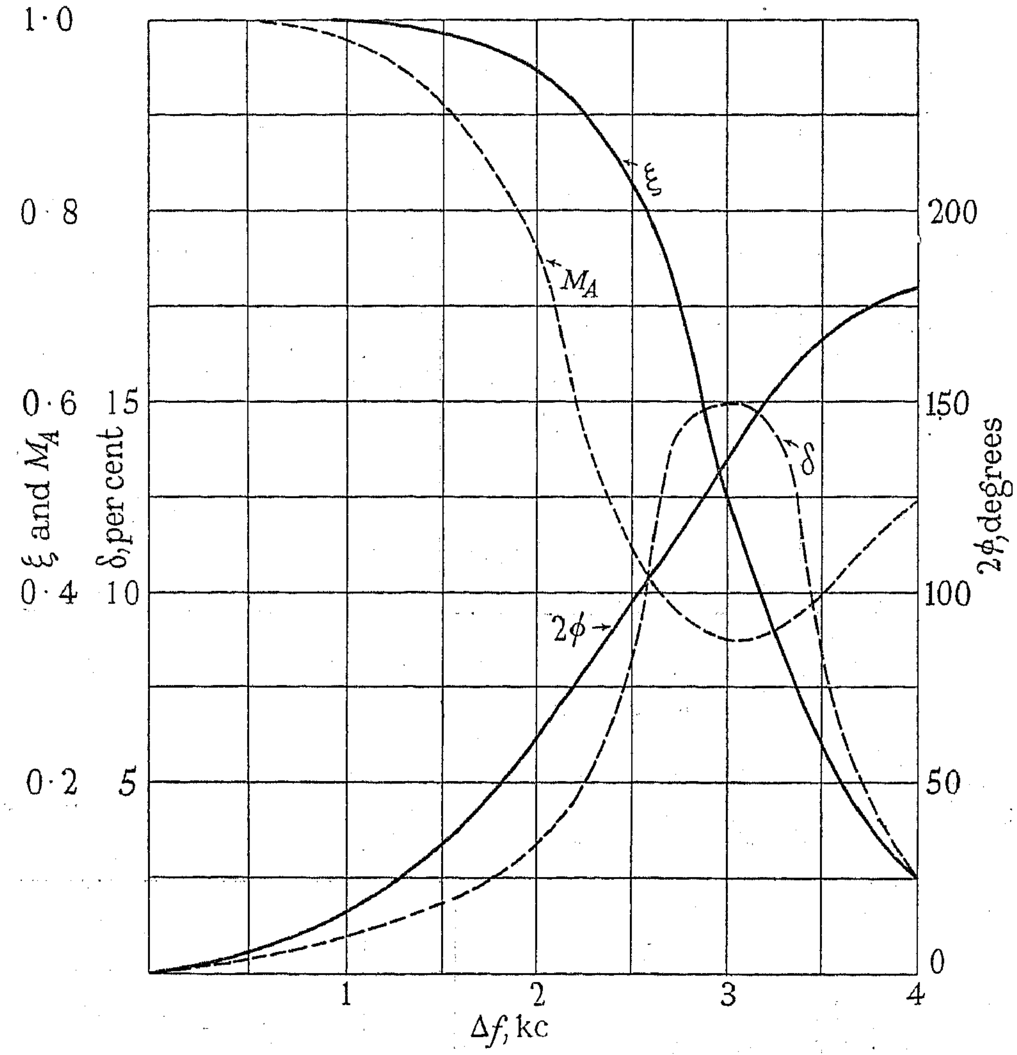

Fig. 21.—Filter comprising constant-k section terminated by f_∞ section. (Experimental values of M_A shown in Fig. 27.)

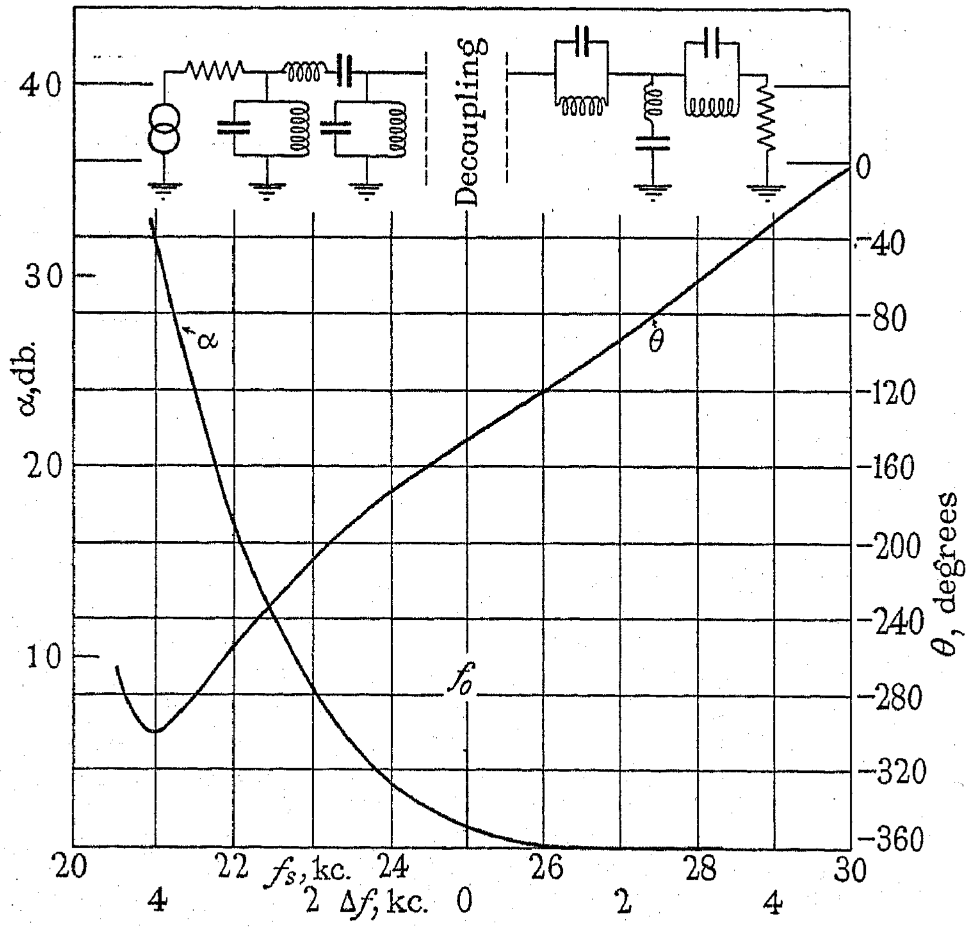

Fig. 22

f_1 (band pass) = 24 kc./sec. f_2 (band eliminator) = 22 kc./sec. f_1 (band eliminator) = 18 kc./sec. Points obtained experimentally.

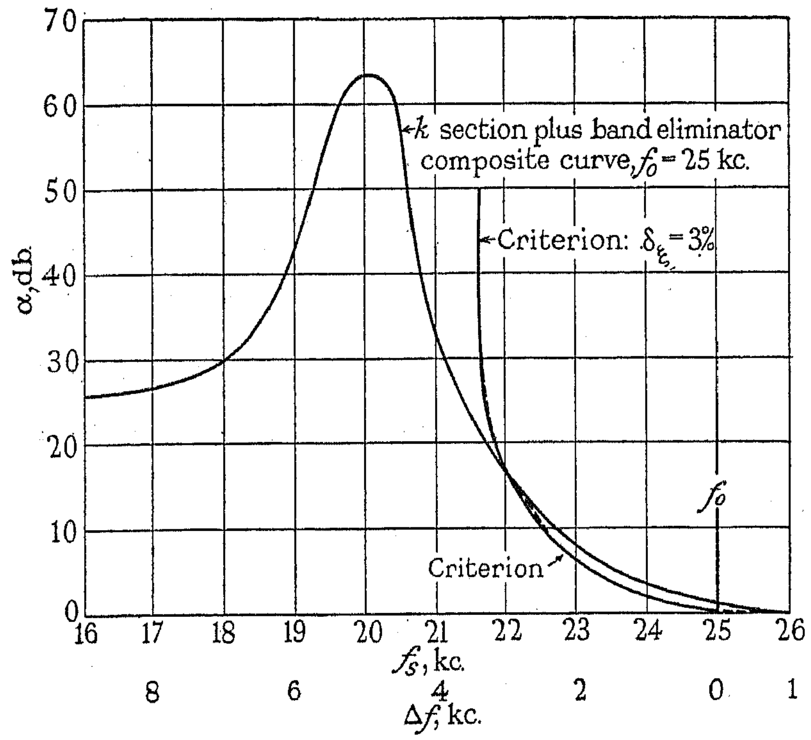

Fig. 23.—Constant-*k* filter plus constant-*k* band eliminator. Points obtained experimentally.

that the network gives an attenuation of the order of 60 db. for sideband components 4–6 kc./sec. removed from the carrier frequency. This is a result which could not be achieved by the cascade connection of constant-*k* type band filter sections, because, although the required attenuation characteristics would be obtained from such networks, the phase characteristics would combine to give such phase asymmetry as to introduce intolerable distortions.

Design of Asymmetric Systems: Conclusions

The quantitative analysis has been used to determine approximate designs. The word " approximate " is used advisedly. Firstly, it is fully appreciated that the sound-spectrum curve is an approximation and is representative of probability rather than being a rigid expression of what are the maximum sound-spectrum intensities regardless of the frequency of their occurrence. Secondly, the charts which show the effects of phase and magnitude asymmetry are based upon an approximate formula. Thirdly, it is impossible to shape filter attenuation/frequency characteristics to exact requirements. It is suggested that the analysis, " inaccurate " in a pedantic sense, has had at least this value—that it has guided a choice of a network which seems almost ideally suited for the purpose for which it was chosen. The test of the value of the system must finally rest with experiment, and Section (5) describes the work which has so far been done in this connection.

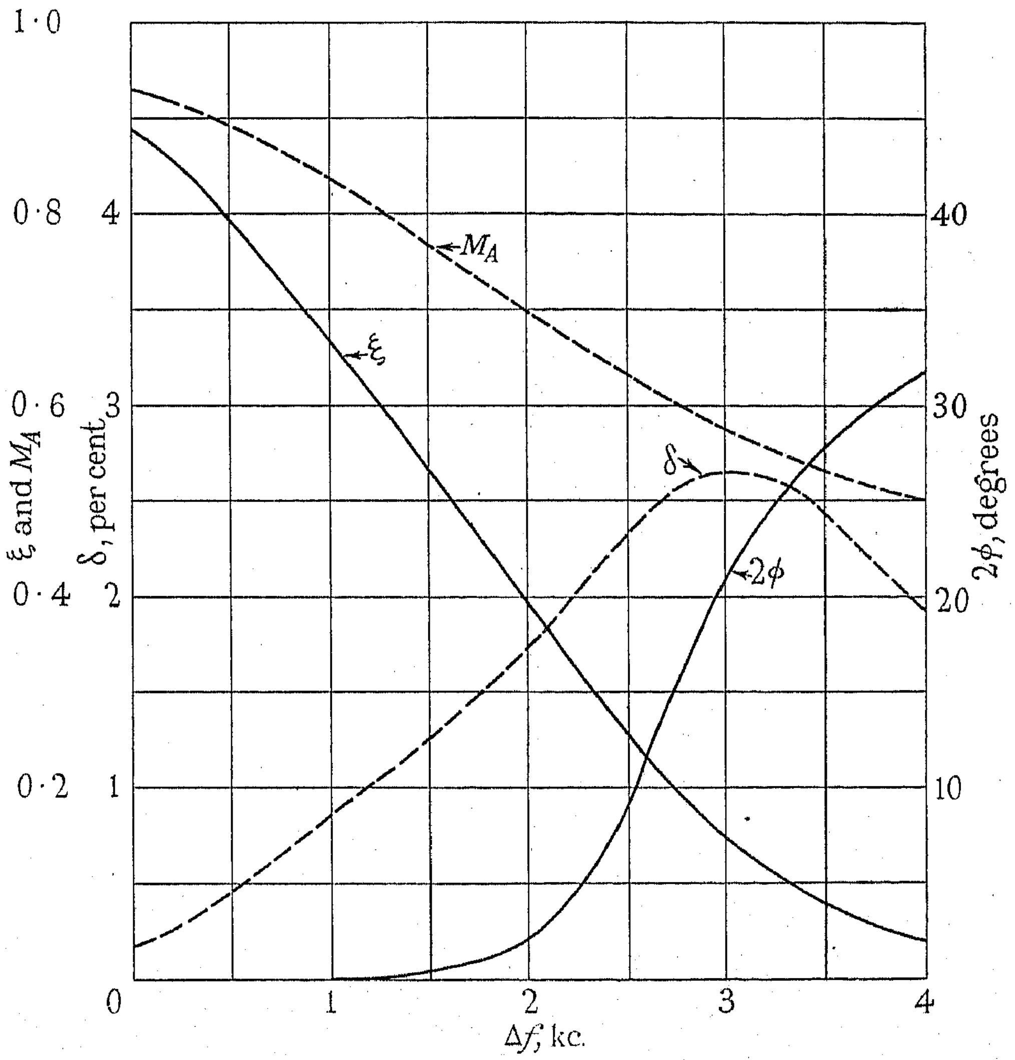

Fig. 24.—Constant-*k* section plus constant-*k* band eliminator.

(5) EXPERIMENTAL VERIFICATION OF THEORY

Receiver Tests

The author's original paper* dealing with the subject of asymmetric-sideband transmission gave accounts of experiments devised to find out, in general terms, whether it was possible to cut off part of one sideband and yet retain reasonably good quality.

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* *Loc. cit.*

A superheterodyne receiver was designed in which the intermediate carrier frequency could be adjusted to lie on various points on the attenuation/frequency characteristic of the intermediate-frequency band filter. It is apparent, now that the analytical work has revealed the way in which distortions may be introduced, that the filter used, a simple 2-section 3-element impedance transforming type, did not introduce phase-asymmetry distortion to any marked degree. On the other hand, it did not cut away as much of the unwanted sideband as has been shown can be removed without the introduction of serious distortion. Correction of the receiver frequency characteristic was made in the audio amplifier stages following the detector.

The quality resulting, even when a considerable portion of one sideband was cut away, was so good that the author felt it worth while to publish his results and invite official consideration of the possibilities of the asymmetric-sideband system.

Transmitter Tests

Test transmitter.

As a result of representations made by the author late in 1936, the B.B.C. supplied funds for the construction of apparatus designed to give an immediate comparison between the highest standard of reproduction possible, according to present-day knowledge, and the quality obtainable from an asymmetric-sideband transmission system.

The apparatus, designed by the author, and constructed in the laboratories of R.M. Radio, Ltd., embodied means for modulating the amplitude of a carrier and routing the modulated currents either through a filter, wherein the carrier frequency was located at f_m, giving symmetrical treatment of sideband components, or through an asymmetric filter. The change-over could be made almost instantaneously by the operation of an external hand switch controlling relays mounted internally on the apparatus panels. The output from each filter was passed alternatively to the same rectifier, which in turn fed its output into an audio amplifier raising the level of the audio component to a value sufficient to operate a rather large high-quality loud-speaker. Other relays, also operated from the external hand switch, served to cut in and out the equalizer necessary to compensate for the demodulating effect of the sideband-cutting filter. Since there is a loss of 6 db. in the equalizer, compensated for by extra magnification in the modulation amplifier, a resistance pad, giving a 6-db. attenuation, was substituted for the equalizer when the symmetric system was in circuit; and so the audio output level, whichever filter was in circuit, remained the same without it being necessary to change the modulation magnifier constants. The carrier frequency, arbitrarily chosen so far as these experiments were concerned, was 27 kc./sec., and the filters were designed, in both cases, to give full reproduction up to modulation frequencies of 10 000 cycles per sec. The steady carrier power was of the order of 1 W. An input modulation level of 1 mW amplified to give a modulation of the carrier of 100 % produced a power of 7–8 W at the loud-speaker terminals.

All conventional circuits for amplification modulation and rectification were designed with a large factor of safety, and, in its final form, either system, symmetric or asymmetric, for frequencies of modulation so low that no sideband asymmetry could occur in the asymmetric filter, gave no more than 2 % 2nd harmonic and negligible 3rd harmonic at the loud-speaker terminals when the carrier was modulated to 90% from an input of 1 mW, the power output into the loud-speaker being of the order of 7 W.

It is clear that everything was done to make the standard quality as pure as possible and that a direct comparison could be instantaneously made between this standard and that obtainable from the asymmetric system.

All the circuits were boxed and panel-mounted.

Test with constant-*k* band-pass section terminated by f_∞ section.

At the time when the first asymmetric filters were designed the author had not appreciated that severe distortions might be produced by phase asymmetry. The theory of amplitude-asymmetry distortion was, however, appreciated and it appeared obvious that the ideal attenuation curve could be most nearly approached by using a constant-*k* section joined to and terminated by an f_∞ section. It has been shown that this combination, from the point of view of the introduction of phase-asymmetry distortion, is, to say the least, an unfortunate choice, but perhaps if the mistake had not been made and its effects observed the solution, now proved so efficacious, would not have been so quickly found.

The results of the experiments made to test the degree of distortion and demodulation caused by the use of the network described give a confirmation of the theory previously set out in this paper.

Fig. 25, already referred to, shows the attenuation and phase characteristics of the network, and Figs. 20 and 21 have previously given the analysis of these characteristics in a form convenient for calculating distortion. An experiment was performed in which the audio-frequency harmonics were measured for a modulation frequency of 3 000 cycles per sec., the carrier being located at 27 kc./sec. The ξ and ϕ values being known for this modulation (and hence sideband), the frequency distortion may be read off from Fig. 11A (in the case considered) and compared with the values found by test. The results are shown in Fig. 26. Fig. 27 shows the frequency characteristic of the asymmetric system without correction, and is again compared with the theoretical values found from the charts of Fig. 12, ξ and ϕ being known from Fig. 21. The agreement is good, considering that the audio-frequency generator used was a rather old model and its calibration not very reliable. Shifting the experimental curve, at about 2 000 cycles per sec., 200 cycles per sec. (8 to 10 %) to the left would give almost perfect agreement. It must further be realized that the phase and attenuation characteristics from which Curve (*b*) in Fig. 27 are calculated were also found experimentally; and so, taking these factors into consideration, one may be very satisfied with the reliability

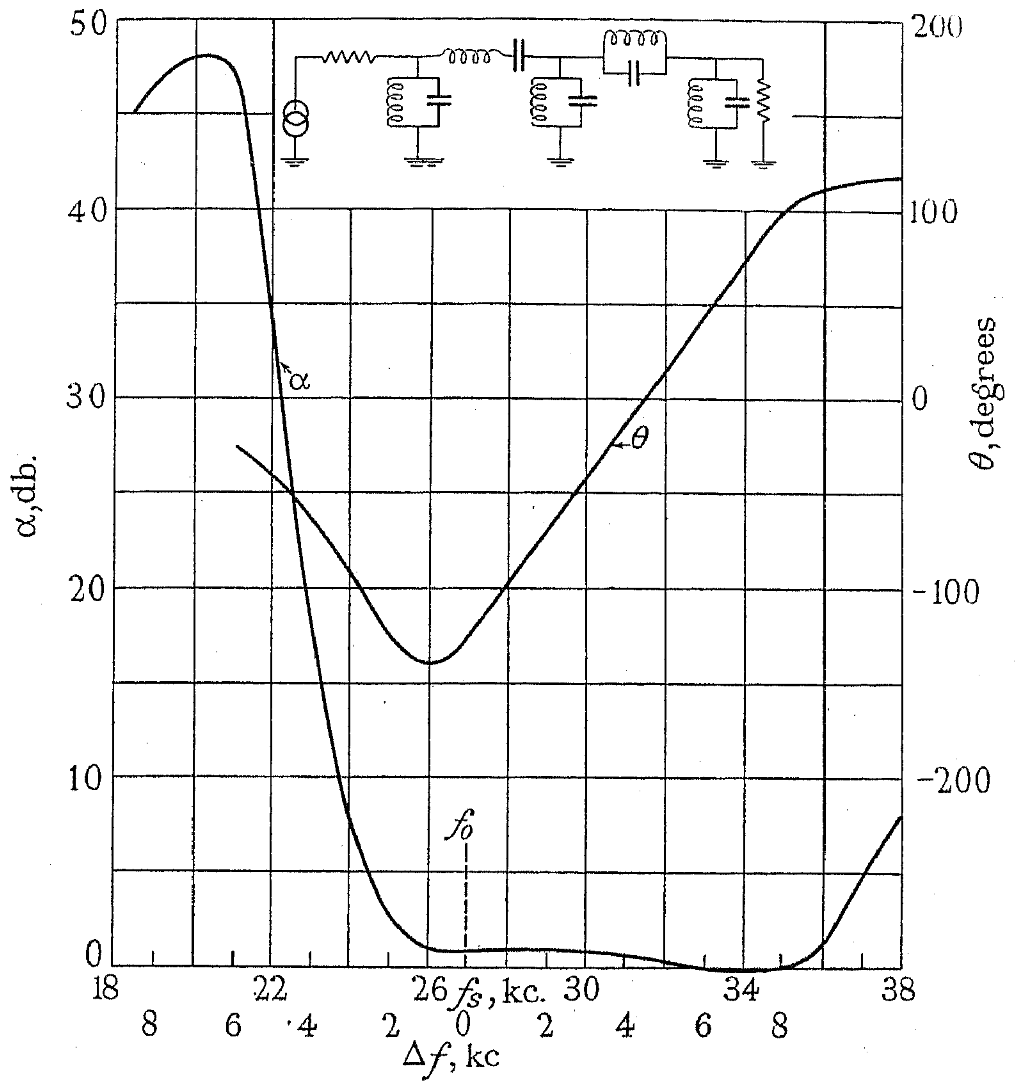

Fig. 25

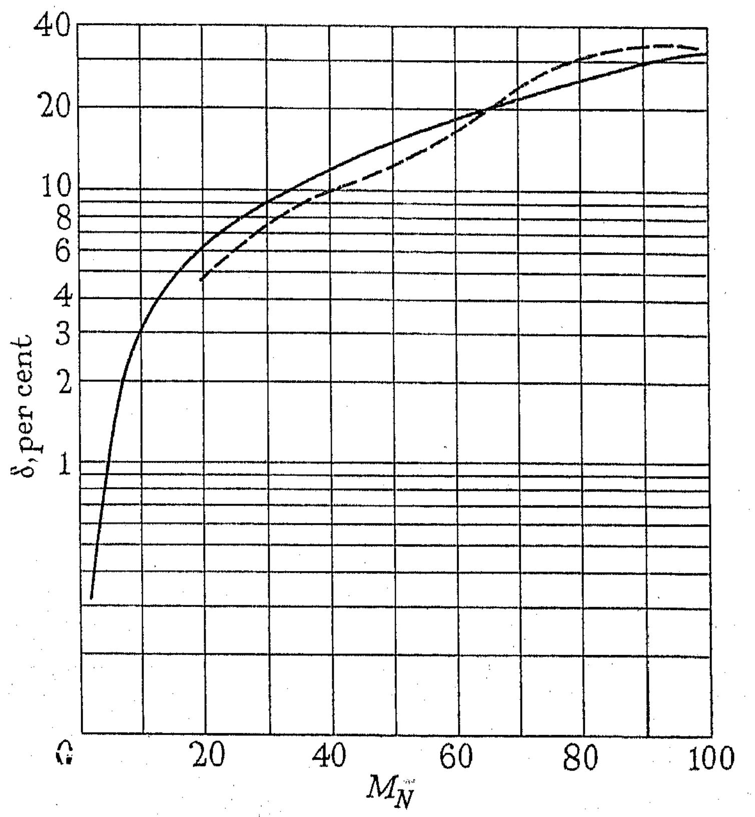

Fig. 26

Modulation frequency = 3 kc./sec. $k + f_\infty$ filter carrier at 27 kc./sec. δ represents percentage distortion of 2nd harmonic.

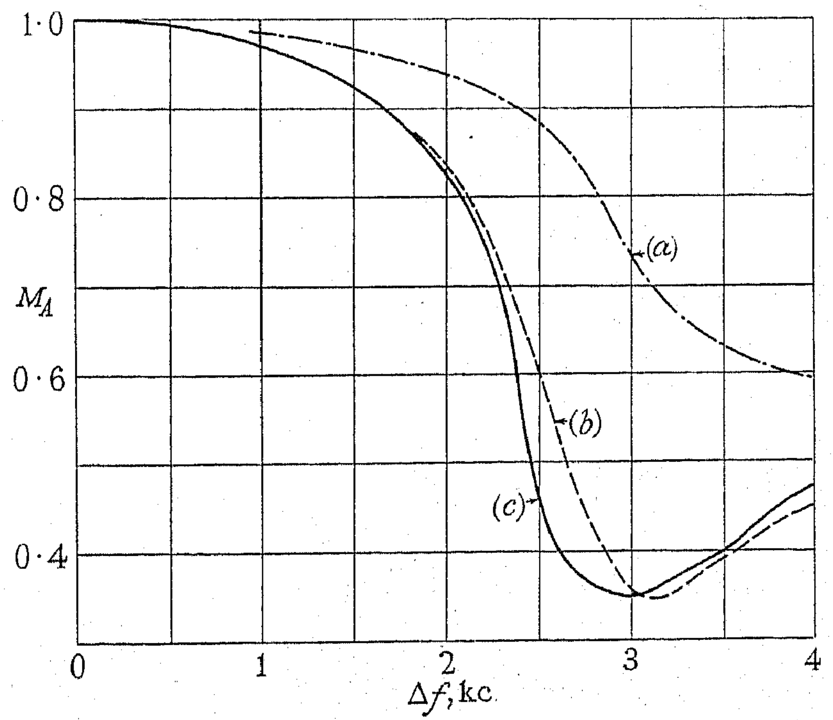

Fig. 27

k filter $+ f_\infty$ section. $f_0 = 27$ kc./sec. (see Figs. 20, 21, and 26).

(a) M_A calculated from $(1 + \xi)/2$, i.e. neglecting phase asymmetry.

(b) M_A calculated from Fig. 12.

(c) Experimental values of M_A.

of the analysis and its usefulness in guiding a choice of circuit values.

Design of equalizer for test transmitter (first design).

In spite of the fact, revealed by the tests, that a fundamental error had been made in neglecting phase asymmetry in the design of the filter, it was decided nevertheless to make comparative tests of quality between the orthodox and the asymmetric system. It was therefore necessary to design an equalizer to correct the demodulation produced by phase and magnitude asymmetry.

It is seen from Fig. 27 that the phase asymmetry demodulates the carrier by more than half and that therefore, as has been shown previously, the equalizer system would have to embody a passive network and a frequency-discriminative amplifier. This would have introduced greater complication in the change-over switching arrangements than was warranted considering the unsuitability of the filter; and so the passive network shown in Fig. 17B was designed in combination with extra amplification to give the best compensation possible in the circumstances.

The overall frequency characteristic of the asymmetric transmitter is given in Fig. 28, showing that the compensation was by no means perfect.

Result of comparative tests of quality (first design).

A receiver containing fixed filters, a high-frequency magnifying valve, and a detector, and designed to give the least possible distortion, was made responsive to the transmission from the London Regional station and its output was fed into the modulation terminals of the test transmitter. This formed the programme material for the comparative tests. The loud-speaker was placed in a room adjoining that in which the change-overs were made, and switching was so arranged that the two reproductions were distinguished by the showing of a green or red light on the loud-speaker baffle. Further switching was arranged so that either the green or the red might apply to either the orthodox or the asymmetric transmission. It was found that a different " click " sound might be heard unless the loud-speaker was short-circuited during the change-over, and the precaution to eliminate this distinctive sound was accordingly taken. There was thus no chance that the observer could be distracted by any differences between the two conditions other than those inherent in the electrical systems employed.

The author does not claim to be able to detect slight differences in quality so easily as can certain rare people. He would be inclined to say that he was unable to detect any difference between the reproduction of the orthodox and of the asymmetric transmissions, except very occasionally, and even then with no real certitude, even though the design of the circuits of the asymmetric transmitter was far from ideal. Observers with ears more acute to distinguish fine differences of quality (two from the B.B.C. Research Department) were therefore asked to judge whether they were able to detect any differences. Several hundreds of change-overs were made during two series, one in the early and the other in the late evening. The programme material, during the first session, combined speech, solo singing, and a light orchestral combination; in the second session dance music formed the bulk of the entertainment. It appeared easier for the observers to distinguish between the two types of transmission during the second session, but this might be explained because modulation is usually deeper during the relaying of dance music. The extreme similarity between the two reproductions may, however, be gauged by recording that the observers, of whom the author was then one, voted, during one set of change-overs, with considerable confidence, for a preference for the quality when one light showed rather than the other, when, in fact, they were listening to the same reproduction, the lights having been changed and not the circuits. It is, however, fully admitted that on balance the observers were more often right than wrong, and that differences could be detected.

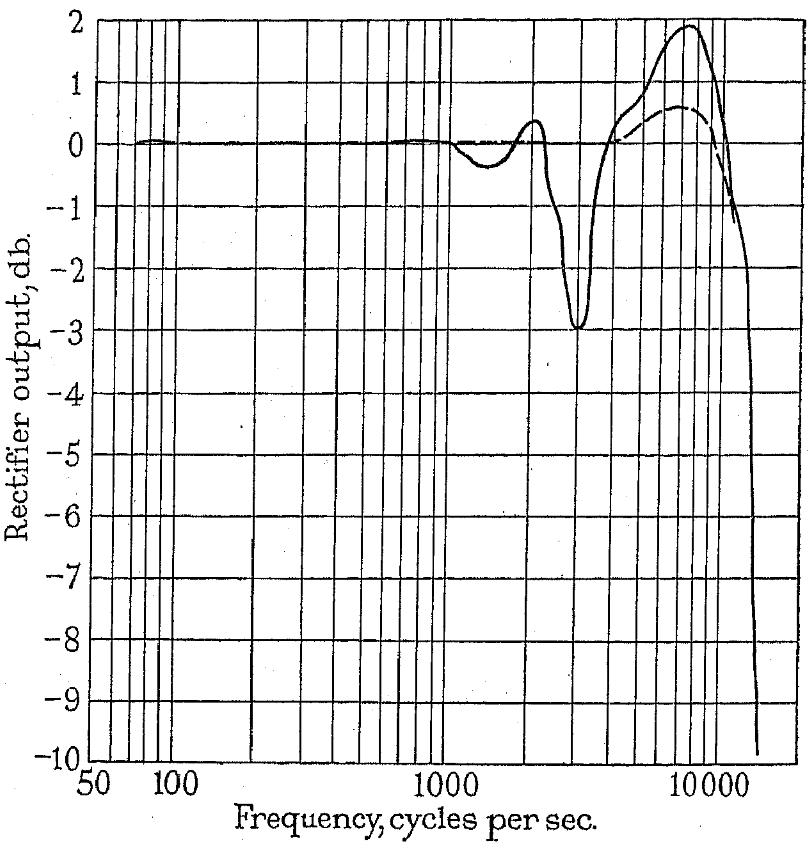

Fig. 28

Constant voltage of modulation. $k + f_\infty$ filter carrier at 27 kc./sec.
——— Output from asymmetric filter with equalizer in.
- - - - - Output from symmetric filter.

Taking 100 change-overs as a basic figure, it is estimated that in 50 cases no differences could be observed or the two observers cancelled each other's preferences, but that out of the remaining 50 the observers were right 35 and wrong 15 times.

This result was very encouraging, considering that it was known by experimental tests that the asymmetric system introduced 30 % distortion at 3 000 cycles per sec. for 100 % modulation (see Fig. 21), showing there would have been maxima of the order of 15 % if equalization had been perfect, but, say, 10 % without the full compensation; and considering that the demodulation effects gave an uneven frequency characteristic (see Fig. 28). The author feels certain that, had frequency multiplication been applied and the two outputs been taken to the input terminals of a typical commercial set, it would have been impossible to distinguish any differences. It is also stated with confidence that none but highly trained observers could have remarked the quality change, because tests were made with a few persons not expert in judging quality and in no case were they able to tell the difference when the circuits were changed.

The results of these first tests were so encouraging that a study was made with a view to devising a circuit which would eliminate the phase distortion but give large attenuation of the unwanted sideband components. The idea of associating the inverse filters was thus conceived, but owing to pressure of other work it was not until the autumn of 1937 that the apparatus was again put into commission and experiments continued.

Tests with constant-*k* band pass and constant-*k* band eliminator (second design).

A filter was built containing the separate networks as described, and the band-pass and band-eliminator circuits were coupled by valves, not resistance networks as shown in Fig. 16. Opportunity was taken to test the findings of the quantitative analysis, and good agreement was proved after certain circuit aberrations had been removed.

Two B.B.C. observers were again asked to make a comparative test of quality between the symmetric and asymmetric systems, the latter being designed, according to the foregoing ideas, to eliminate phase asymmetry. Extremely exhaustive tests were undertaken to make sure that the symmetric system was, so far as is humanly possible, quite free from distortion. These tests included an examination of the wireless receiver which provided the programme input. It was proved that the system, from aerial input to loud-speaker output, gave no more than 1·5 % 2nd-harmonic and negligible 3rd-harmonic distortion. The frequency characteristics of the two transmitters, symmetric and asymmetric, were substantially the same, the maximum departures being of the order 1 db. at frequencies above 6 000 cycles per sec. The bass response was 0·6 db. less for the asymmetric system at 50 cycles, and this difference decreased until it was zero between 1 000 and 6 000 cycles.

The tests themselves were perforce of limited duration and were to a point, therefore, inconclusive. It has been decided that far more exhaustive tests must be carried out in the future.

Nevertheless, the following conclusions are fairly definite.

(1) It is agreed that the system wherein the band-pass and band-eliminator networks are used in combination is a great deal less liable to give audible distortion than that in which an f_∞ section is associated with the constant-*k* filter. Thus in the previous tests the observers were able to detect the difference in quality between symmetric and asymmetric transmission with some degree of certainty, given certain conditions of modulation and programme item. In the tests on the new transmitter the observers were, over a limited period, more often wrong than right in their choice, believing the asymmetric system to be actually more pleasing and less "rough" than the symmetric.

(2) The observers remarked that they thought the quality on both transmissions inferior and that this fact was inclined to vitiate any firm conclusion. The author feels that, while this may conceivably have been due to some fault in the test apparatus or, more probably, in the loud-speaker used, nevertheless the very critical analysis of distortion and frequency characteristics would suggest that in this instance the source itself was not up to standard.

(3) The most remarkable fact adduced was that the observers were inclined to distinguish differences of frequency characteristic between the transmissions, e.g. there appeared to them to be more bass in one than the other. This, in view of the measurements, is very surprising although a quite definitely established fact. The only conclusion one may draw is that a decibel of difference is distinguishable by acute ears, provided one supposes that this difference extends over a fairly wide frequency-range.

The author, in any case, feels justified in believing that the solution of the problems of obtaining a very large attenuation of one sideband without the introduction of phase-asymmetry distortion has been achieved. Whether or not amplitude asymmetry introduces detectable distortion remains to be proved, but first tests indicate that it is far from easy, given even the most acute ears and a standard for comparison, to do so.

Another experiment, relevant to the following Section which deals with the advantages of asymmetric transmission in eliminating "sideband splash" or spectrum-overlap interference, was also demonstrated. In this experiment the modulated carrier-frequency output from the test transmitter having a carrier frequency of 25 kc./sec. was applied to a frequency multiplier, and the multiplied frequency of 250 kc./sec. was fed in turn into a typical commercial wireless receiver. An unmodulated carrier, having a difference frequency from the multiplied output from the test transmitter of 9 kc., was also fed into the receiver. The receiver was then tuned to the unmodulated carrier and conditions were arranged so that typical sideband interference caused by the test transmitter working on symmetric modulation was heard. Evidently the unmodulated carrier represented a wanted station while the output from the test transmitter was typically that of a powerful unwanted station on the frequency-contiguous channel, causing the well-known "sideband splash" interference.

Pressing the button which changed the output of the test transmitter from symmetric to asymmetric transmission greatly changed the degree of interference heard, because the sideband causing the interference was (partly) cut away. It was estimated, by the ear, obviously a very crude instrument for measuring effects quantitatively, that the interference was reduced tenfold. It was not, however, wholly eliminated, nor, as will be seen from theory, could it have been with only a 9-kc. separation and a modulation up to 10 000 cycles.

Lastly, means were arranged to create a background noise, and when this contained high-frequency audio-components the change-over to asymmetric transmission showed a considerable amelioration of the noise, particularly as regards its high-frequency components; demonstrating that, in the asymmetric system, the signal/noise ratio is superior to that obtainable with symmetric transmission.

Great advantages are seen to accrue from the cutting-away of part of one sideband, and these, in the author's opinion, outweighed the introduction of so negligible a harmonic distortion.

(6) REQUIRED CARRIER DIFFERENCE FREQUENCIES

General

The stated object of the preceding analysis was to find the answer to certain questions. It was required to find, *inter alia*, what economies could be effected in the frequency band width normally occupied by the radio-frequency spectrum of a broadcasting transmission, assuming that negligible harmonic distortion was introduced by removing part of a sideband. The analysis might have proved that so little of the sideband could be cut away without introducing more distortion than could be tolerated that it was not worth while to make any changes to existing technique; on the other hand, theory might have shown that by far the greater part of the sideband could be removed while still maintaining good-quality reproduction, thereby proving the advisability of using the system. This Section is therefore devoted to finding the required frequency-separation between the carrier frequencies of broadcasting stations, when different types of transmission technique are employed. It is necessary to make a classification of the variables so that the conclusions may not be ambiguous.

There is an obvious distinction between two methods of expressing the advantages of the asymmetric system. Thus we may either make a study to find out what is the minimum frequency-separation between carriers to ensure that, in no circumstances, can the radiated spectra overlap (and thus make sources of potential interference); or we can fix upon some frequency-difference established by *force majeure*, and find out what diminution of inevitable interference will be brought about if the asymmetric system is used.

This distinctive classification can be related either to a broadcasting system which is set up *de novo* and where the choice of carrier frequencies is uninfluenced by political and economic factors which have little to do with purely technical criteria, or to an existing system of broadcasting so established that it would be impracticable to suggest any diminution of the number of working stations, even though this diminution would be far less, using the asymmetric system, than that necessary to give complete freedom from spectrum overlap if double-sideband transmission were employed.

In sum, it is necessary to find: (1) What economies in the use of channels can be made with the asymmetric system applied to any new system which does not impose any other criteria than those demanding good quality and freedom from inter-channel interference. (2) What diminution of inevitable inter-station interference can be expected if the asymmetric system has to be applied to an existing set-up wherein carrier-frequency separations are less than those demanded by the criteria given in (1), i.e. when it is an essential of the application of the system that the existing service conditions shall not be disturbed. There is, lastly, a further subdivision of (2) which is concerned with the type of transmitter at present existing and whether or not it may be economically and practically adapted to make full use of the asymmetric principle. It will be shown in this connection that, if modulation of the carrier takes place at a level approximating to its maximum level (high-power modulation), it is probable that the ideal network, consisting of a constant-*k* band filter and a constant-*k* band eliminator in conjunction, cannot be used. If this be the case we could not expect a better result than that given by a constant-*k* band filter alone. If the existing transmitter employs low-power modulation, wherein, as the term implies, modulation of the carrier takes place at a level greatly below the radiated carrier level, modification is comparatively simple and the full advantages of the ideal network may be counted upon.

Therefore, in studying the effects of the introduction of the asymmetric system and attempting to assess its benefits, we shall make a classification of the possible variations as follows:—

Classification (1). Minimum carrier difference frequency to prevent spectrum overlap.

Classification (2). The degree of interference experienced with a given carrier difference frequency. Thus we must find:—

(1) (A). Minimum carrier-wave separation, assuming the use of double-sideband transmission.

(1) (B). Minimum carrier-wave separation, assuming the use of asymmetric transmission, and a constant-*k* band-pass section plus a constant-*k* band-eliminator section with demodulation correction.

(1) (C). Minimum carrier-wave separation, assuming the use of asymmetric transmission, a constant-*k* section only, and demodulation correction.

(1) (D). Minimum carrier-wave separation, assuming the use of asymmetric transmission, a constant-*k* band-pass section plus a constant-*k* band-eliminator section, and no demodulation correction.

Dealing with the last four items, (1) (A) applies to existing conditions; (1) (B) applies to existing space broadcasting transmitters having low-power modulation, but assumes that existing carrier differences can be increased (probably an impracticable suggestion); (1) (C) applies to existing space broadcasting transmitters having high-power modulation, but assumes that existing carrier differences can be increased (probably an impracticable suggestion); (1) (D) is theoretical.

(2) (A). Amount of interference expected in existing conditions (9-kc. carrier-wave separation).

(2) (B). Amount of interference expected using asymmetric transmission, a constant-*k* band-pass section plus constant-*k* band eliminator, and demodulation correction (9-kc. carrier-wave separation).

(2) (C). Amount of interference expected using asymmetric transmission, constant-*k* band-pass section only, and demodulation correction (9-kc. carrier-wave separation).

(2) (D). Amount of interference expected using asymmetric transmission, a constant-*k* band-pass filter plus a constant-*k* band-eliminator section, and no demodulation correction (9-kc. carrier-wave separation).

It should be remarked that (2) (A) applies to the existing condition; (2) (B) applies to the condition which will apply if existing carrier-wave separations are maintained but the low-power-modulation transmitters are adapted; (2) (C) is the same as (2) (B), except that it applies to existing high-power transmitters; (2) (D) is theoretical.

Method of Analysis

A description of the method by which the required frequency-separation between two carriers surrounded by any spectrum can be calculated was first published by the author some 5 years ago.* Its re-statement here is not redundant, partly because the results obtained are a vital part of the analysis of the asymmetric-sideband system, and partly because it may be possible to give a clearer picture of the implications of the findings than was possible 5 years ago.

The method involves first the determination of boundaries enclosing an area, having the vertical dimensions of power level and horizontal dimensions of radio frequency, within which it may be said that the spectrum energy of a modulated carrier has " appreciable " magnitude and is sufficiently spread in a horizontal dimension to allow a reproducer energized by the full spectrum to give a required fidelity of reproduction. The carrier is represented by a vertical line the top of which coincides with what will be arbitrarily called zero level. In a double-sideband transmission the maximum level (corresponding to 100 % modulation) of the sideband components will be 6 db. below zero (or carrier) level. Throughout this paper 80 % modulation is chosen as a maximum, and so in a symmetric system the maximum sideband level is 8 db. below zero level. Accepting the premise that the sound spectrum intensities are distributed as indicated in Fig. 14, then the maximum sideband level of 8 db. will occur at radio frequencies of $(f_0 \pm 0{\cdot}2)$ kc./sec. to $(f_0 \pm 1{\cdot}2)$ kc./sec. The maximum levels of sideband components created by modulation frequencies greater than 1 200 cycles per sec. will be relatively distributed according to the power ratios corresponding to the modulation factors given in Fig. 14. Thus two upper boundaries may be drawn symmetrically about the carrier representing the sideband, and hence the radio-frequency spectrum, maxima. Part of one and all of the other side of the spectrum created by symmetric-sideband transmission is shown by the part 1A in Fig. 29(1).

It is now necessary to find a lower boundary to delineate the levels at which it may be said the spectrum has " appreciable " intensity. Assume that the spectrum is received rectified and made to energize a loud-speaker, that no distortion whatsoever is introduced in the process, and that the receiver is perfectly selective and will not appreciate energy lying outside the frequency limits occupied by the spectrum. The spectrum-intensity variations are then evidently converted to audio-frequency variations, which, in turn, are made to create, through the agency of a loud-speaker, pressure variations in the air which are detected by the ear as audible sound. Presumably the receiver will be so adjusted that the maximum sounds (corresponding to the condition when the sideband components in the spectrum touch the boundary level drawn as described) are of a level to satisfy the tastes of the listener. Now if any unwanted or parasitical wave energy, having frequencies which lie between the frequency limits occupied by what we may call the " wanted stations spectrum," had a sufficient intensity, it would create audible and unwanted sounds in the loud-speaker output. On the other hand, if the intensity of the interfering-wave energy was less than a certain amount it would not create audible interference. Presumably, however, assuming a " perfect " reception system, it would create some corresponding air-wave energy, however feeble. The only basic reason, therefore, why very feeble interference is not heard is because the human ear itself is incapable of detecting it. This means that the lower boundary of the area within which energy levels may be described as " appreciable " is the ear threshold-of-hearing curve. If the ear were infinitely sensitive, then, apart from the effect of the masking of small sounds by the louder, it would appreciate in the postulated system the effects of all noise of whatever level. A " perfect " crystal set tuned to a local station is in one sense very selective but only because the ear is, by its limited sensitivity, also selective.

It is now necessary to choose the relative levels between the minimum of the ear threshold-of-hearing boundary and the maximum sideband intensity so that a complete area may be enclosed within which we may say wave energy can be appreciated in ideal conditions of reception.

The choice of this (decibel) scalar distance is a matter of the choice of the " contrast " level necessary to give as faithful a rendering of the original as possible. If we attempt to imitate too nearly the conditions of the concert hall, background noise becomes too loud to be eliminated by any but the strongest signals. It is said that a symphony orchestra has a contrast level between *ppp* and *fff* of 80–90 db. The ear has a range between sensation level and ear threshold-of-hearing of 130 db. Allowing for the many difficulties inherent in the design of high-fidelity systems, and allowing for a reasonable signal/noise ratio, it is proposed to choose 60 db. as a contrast level desirable for good-quality reproduction. Choosing this value permits the area to be enclosed as shown by the part 1A in Fig. 29(1).

It will be clear that if reception is to remain, in all circumstances, unspoiled by unwanted interference, then the area must be considered sacrosanct and must never be invaded by energy components created by unwanted stations or by sources of parasitical waves. We shall only consider hereafter the possibilities of invasions of energy due to spectrum components created by other transmissions located on frequency-contiguous channels, i.e. we shall not consider questions of " noise " invasions.

The method of determining the minimum required carrier difference frequency can now be more fully explained. We distinguish first between a wanted station, always labelled " W," and an unwanted station, distinguished on the diagram by the letter " U." A shape is cut out, preferably from thin celluloid, which is the shape of the sacrosanct wanted-stations frequency-decibel area that must never be invaded if interference is to be negligible. Another shape is similarly prepared, the upper limits of which are drawn to represent the carrier and sideband intensities and therefore the upper limits of the energy created by an unwanted station's transmission. The combination of cut-offs introduced in the apparatus of transmission and the fact that the sound energy itself is so relatively feeble at high modulation frequencies allows the upper boundary to become substantially a vertical line at carrier difference frequencies of 8 000 cycles per sec. But the area representing the invading energy extends theoretically infinitely downwards from zero level. Since we are only interested in its value in relation to a limited vertical scale, it is not necessary, except in certain cases where one sideband is attenuated, to consider any limits to the lower boundary. Thus in the part 1A in Fig. 29(1) the unwanted station's energy is only delineated as regards its maxima.

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* " The Required Minimum Frequency Separation between Carrier Waves of Broadcast Stations," *Proceedings of the Institute of Radio Engineers*, 1933, vol. 21, p. 193.

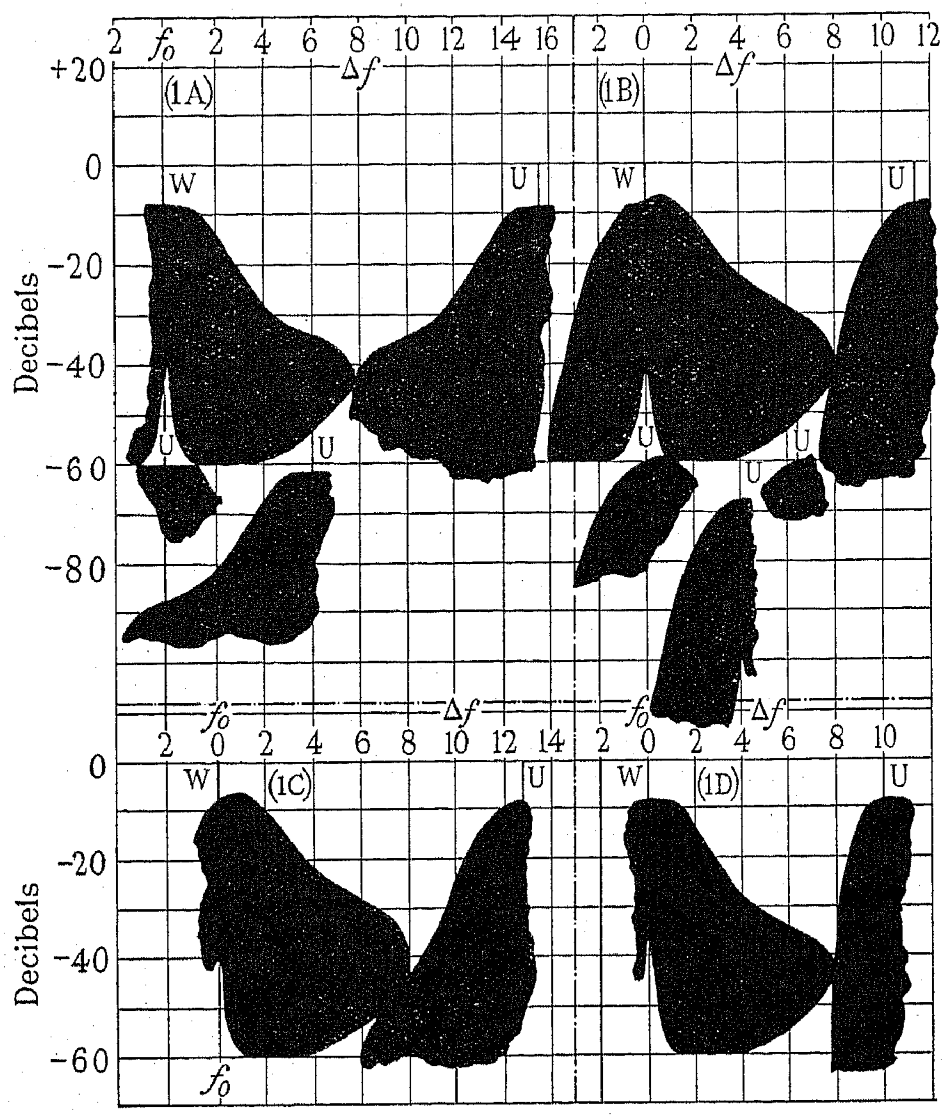

Fig. 29(1).—Minimum required carrier difference frequencies.

1A: Double sideband. 1B: Asymmetric; *k* filter plus band eliminator, demodulation correction. 1C: Asymmetric; *k* filter only, demodulation correction. 1D: Asymmetric; *k* filter plus band eliminator, no demodulation correction.

The wanted-station shape is anchored on scaled squared paper with its carrier components vertical; and the unwanted-station shape is held, also with carrier vertical, in various relative positions, it being always ensured, however, that the shapes shall not overlap. The horizontal scalar distance between the lines representing the carriers is the frequency separation, and the vertical scalar distance between the carrier tips is the relative level. This indicates the required minimum carrier difference frequency for interference-free reception, given different relative fields at any one reception location.

If now the wanted-station area is again fixed but the unwanted is separated by sub-optimum carrier difference frequencies, then there is an area of overlap. This area, varying with a change of relative levels, is a measure of the interference produced when the carrier-frequency separation is smaller than that necessary to prevent interference. The determination of the conditions can be performed with different shapes representing the different spectra created by various types of transmission.

Preparation of Required Spectrum Areas

Double sideband.

In Fig. 29(1) the part 1A shows the required spectrum areas for double-sideband transmission and has been prepared according to methods previously described. The lower-frequency sideband spectrum of an unwanted station is also shown occupying the next higher-frequency channel to that occupied by the wanted station, which has its upper-frequency sideband fully represented. The two transmissions are of equal level and their spectra touch but do not overlap.

Asymmetric.

The parts 1B, 1C, and 1D in Fig. 29(1) show different asymmetric spectra. The conditions are given in the Figure, and the figures and letters chosen are in accordance with the schedule of conditions previously set out under " General " in this Section. The following is the method of preparation of the required shapes.

For the fully transmitted sideband: If no demodulation correction is supposed to take place then the fully transmitted sideband spectrum shape is the same as that representing the double-sideband case. If, however, correction is introduced, obviously the level corresponding to that of the sideband maxima when double-sideband transmission is represented is multiplied by $1/M_A$ (see Fig. 12).

For the attenuated sideband: This is first drawn as for the fully transmitted sideband. The decibels of attenuation given by the curve representing the attenuation/frequency characteristic for the circuit supposed to be used are then subtracted from the curve drawn to represent the fully transmitted sideband. The resulting boundary gives the maxima of the attenuated sideband.

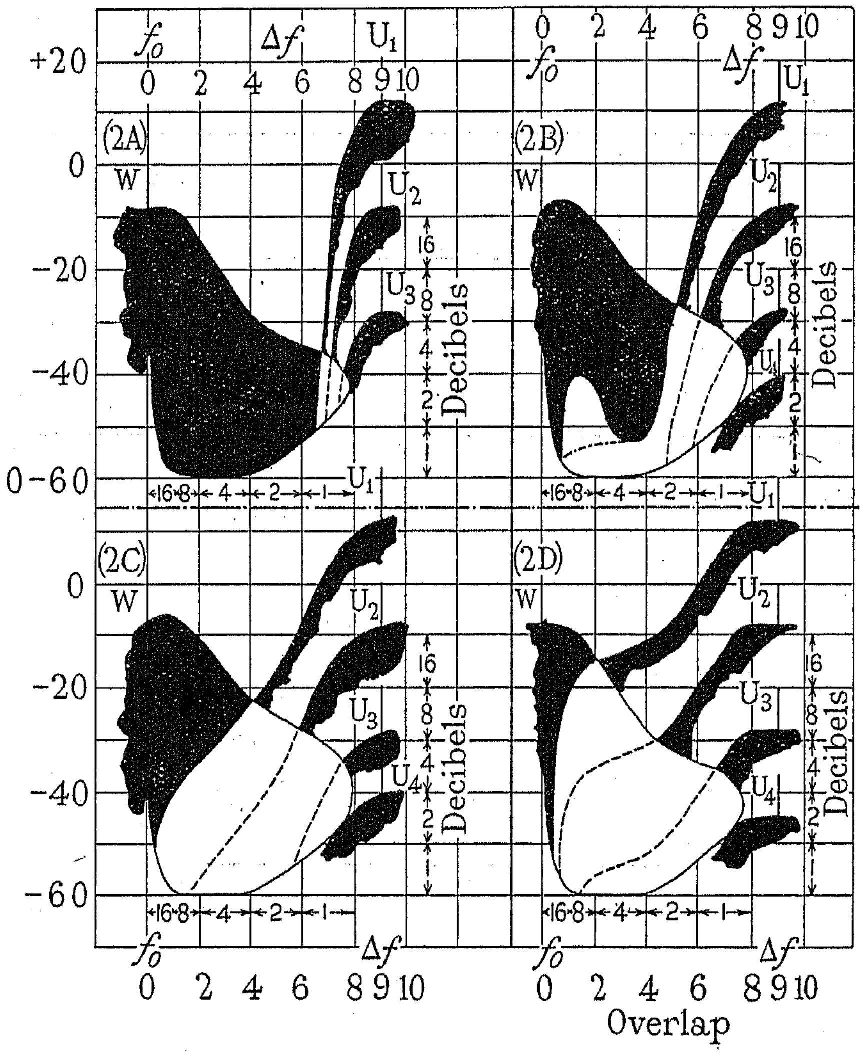

Fig. 29(2).—Spectrum-overlap interference.

2A: Asymmetric; *k* filter plus *k* eliminator, no demodulation correction.
2B: Asymmetric; *k* filter plus *k* eliminator, demodulation correction.
2C: Asymmetric; *k* filter only, demodulation correction.
2D: Double sideband.

Explanation of Fig. 29(1)

Minimum separation.

The black areas in the parts 1A, 1B, 1C, and 1D in Fig. 29(1) represent some typical positions of the sliding unwanted-station spectrum, but in order to economize space neither all the positions that may be used nor the full extent of the spectra are shown. The conclusions given hereafter were obtained by measuring the relative positions of the various shapes representing the spectra; the diagrams only exist to show how these were typically placed.

Explanation of Fig. 29(2)

Interference.

The degree of interference is assessed in the parts 2A, 2B, and 2C, in Fig. 29(2), representing conditions corresponding to the chosen labelling. The degree of interference must be a function of the amount of the wanted-station's frequency-decibel area which is occupied by the energy components of the frequency-contiguous unwanted station. But the interpretation of the results is not made in terms of the direct ratio of the areas invaded (*a*) when double-sideband, (*b*) when asymmetric technique is used, because this would not give a true assessment of the economies effected by cutting away part of one sideband.

Thus the listener adjusts his receiver, when sideband interference takes place, to cut off the upper parts of the spectrum in order to get rid of extraneous sounds. This process becomes increasingly destructive to good-quality reception as more and more " top " is cut off. If conditions are such that the sideband invasion becomes very pronounced, the listener, rather than forgo all intelligibility, tolerates, albeit grudgingly, some interference. This means that the onslaught of unwanted energy which brings the invading frequency close to the carrier is far more dangerous to peaceful reception than one that is staged in the remoter outer regions of the sideband.

For these reasons the amount of interference is more truly assessed by giving in the invaded areas different " weight " factors than by taking their direct ratio as a measure of annoyance. These weight factors have been chosen as follows: between 8 000 and 6 000 cycles per sec., 1; between 6 000 and 4 000, 2; between 4 000 and 2 000, 4; between 2 000 and 1 000, 8; between 1 000 and 0, 16. Similarly, weight factors are chosen for the vertical scale. These are shown in Fig. 29(2). The black areas are those in which the wanted spectrum is never invaded, while the conditions for different relative levels between wanted and unwanted stations are shown by dotted lines in the white spaces.

It will be remarked that in the part 2B in Fig. 29(2) the attenuation-curve shape using the band eliminator is such that a larger part is carved out of the wanted spectrum than would be the case if an f_∞-section band eliminator were used. The condition has not been studied; it might be that phase-asymmetry distortion would become serious if a more complex network were employed, but this seems unlikely since the tendency to inverse flexures of the phase characteristics of the inverse types of filter sections would still exist.

Increase of Available Channels

Let us assume that it may be argued that, after all, existing interferences are not serious and that the great majority of people listen to the local station and are not bothered as much by spectrum-overlap effects as an inspection of the diagrams would lead the reader to believe. Taking what is, to the author's way of thinking, this false premise as a basis, it is interesting to work out how much closer stations could be moved, if the asymmetric technique were adopted, to give the same degree of interference as exists to-day. Choosing the condition of equal level and the asymmetric conditions 2D and 2B in Fig. 29(2), we find that a separation of 6 kc. could be tolerated, giving an increase of about 30 % in the number of available channels.

Summary of Results

The results are summarized in Table 3.

In considering the "gains" we take, as a criterion, carrier and single-sideband working. It is clear that in a 1 000-kc. gamut, the maximum modulation frequency transmitted being 7 800 cycles per sec., this system would allow 128 free channels. A figure of merit is given to each system, this figure being the ratio of the free channels available for that system assuming no spectrum overlap. The percentage gain is the number of extra channels over and above those given by the double-sideband system divided by the number given by the carrier and single-sideband system, multiplied by 100. The interference figure of merit is the interference factor of a given system divided by the interference factor of the double-sideband system, multiplied by 100.

Table 3*

Condition	Reference in Fig. 29(1) or 29(2)	Practical case	Minimum separation of carriers to prevent overlap (kc.)	Gain of asymmetric over double-sideband in number of available channels for no overlap			Ratio of interference in given system to that in double-sideband system with 9-kc. carrier separation: equal fields at point of reception
				Number of channels per 1 000 kc.	Figure of merit	Gain (%)	
Double sideband	1A	Existing conditions	15·6	64	0·5	—	—
Asymmetric filter and eliminator, demodulation correction	1B, 2B	Results obtainable by modifying existing low-power-modulated transmitters	11·4	88	0·686	19	0·12
Asymmetric filter only, demodulation correction	1C, 2C	Results obtainable by modifying existing high-power-modulated transmitters	12·8	78	0·61	11·9	0·234
Asymmetric filter and eliminator, no demodulation	1D, 2A	Theoretical	10·0	100	0·78	28·1	0·011
Carrier and single sideband . .	—	An eventual possibility	7·8	128	1·00	50	0·00

* If, with the ratio of the fields of the wanted and unwanted stations equal to unity, it was decided that we could support the same degree of interference as is suffered to-day [see 1A, Fig. 29(1)], then the carrier-wave separation for Condition 2B would be 6 kc., giving 150 channels in a 1 000-kc. band, an increase of 40 channels over and above those available with a 9-kc. separation.

Table 3 is supplemented by Figs. 30 and 31, which show some of the information in curve form.

Summarizing, the investigation proves that the asymmetric system makes for great economy in the use of any available frequency gamut. Its application would be a great help in minimizing the all-too-common sideband interferences set up because of the insufficient carrier-wave separations commonly employed. A point of further interest not brought out quantitatively in the analysis is that the signal/noise ratio is increased by the adoption of the system, assuming, of course, that the receiver has also a narrowed band width of response enabling it to benefit from the asymmetric transmission.

(7) RECEPTION PROBLEMS

General

The preceding analysis has assumed that the receiver introduces no distortion and is sufficiently selective to be quite unaffected by any wave energy lying outside the boundaries which enclose an area within which, it has been postulated, wave energy is "appreciable." This Section treats the problems of reception in terms of actualities: no receiver is perfect in the sense postulated, and its departure from perfection must to some extent modify the conclusions.

It is suggested, however, at the outset that it is hardly the business of the transmission authorities to attempt to solve reception problems introduced by changes in transmission technique, provided they can be satisfied that it is not in principle impossible or impracticable to design the circuits required to make full use of the changes. Transmission is to reception what environment is to the evolution of species; it is "up to" the species to modify themselves to the changed environment. It is, however, the business of the transmission authorities to ensure as far as possible that the spectra of stations occupying contiguous channels shall either not overlap or that the overlap is small. If this is not done then the receiver, even if it has an asymmetric response in relation to its response to the carrier component, cannot *per se* eliminate inter-station interference. Thus if a wanted station lies between two neighbours both of which create fields comparable with that of the wanted station at the point of reception, both sidebands of the wanted station will be equally interfered with and single-sideband or asymmetric-sideband *reception* cannot eliminate such interference. On the other hand, it would be of no use in reducing inter-station interference if, assuming asymmetric transmission, the receiver gave a symmetrical response about the carrier and of sufficient frequency-

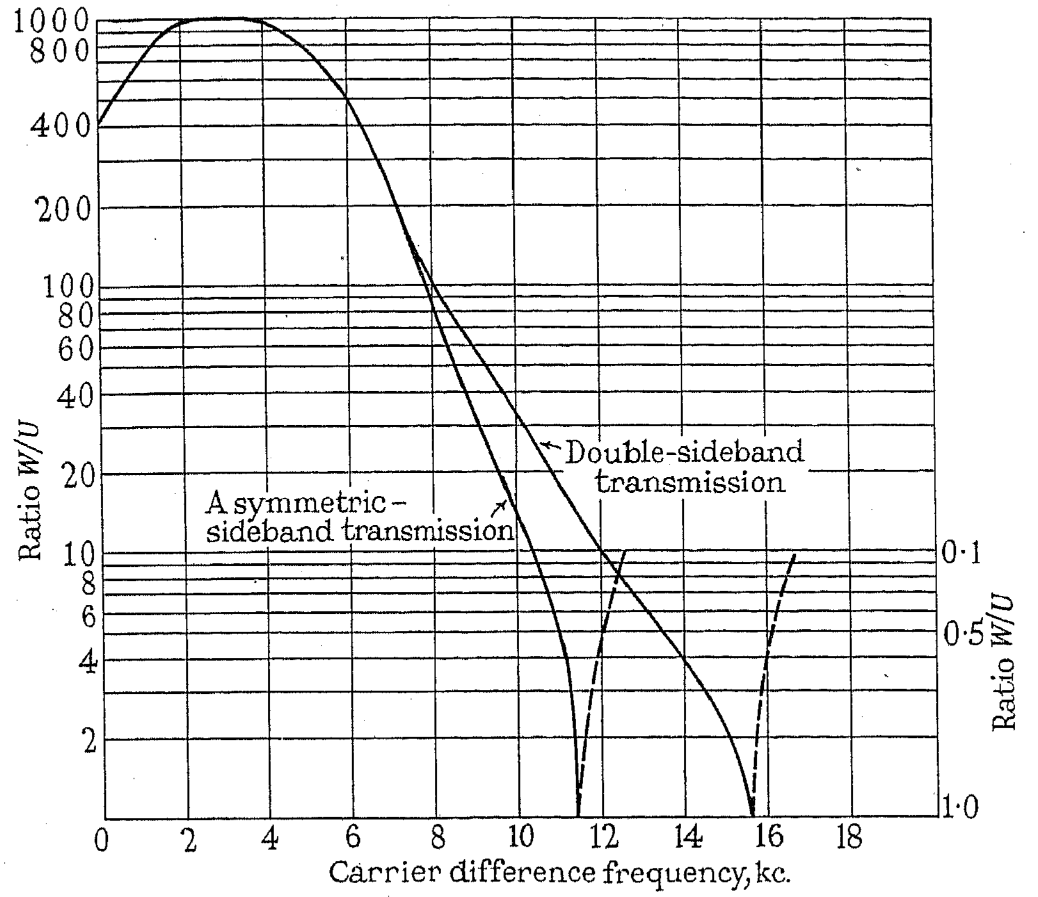

Fig. 30.—Required frequency-separation for no spectrum overlap.

Asymmetric demodulation corrected in transmitter. Contrast level, 60 db. W = wanted station, U = unwanted station. Right-hand scale refers to broken-line curves.

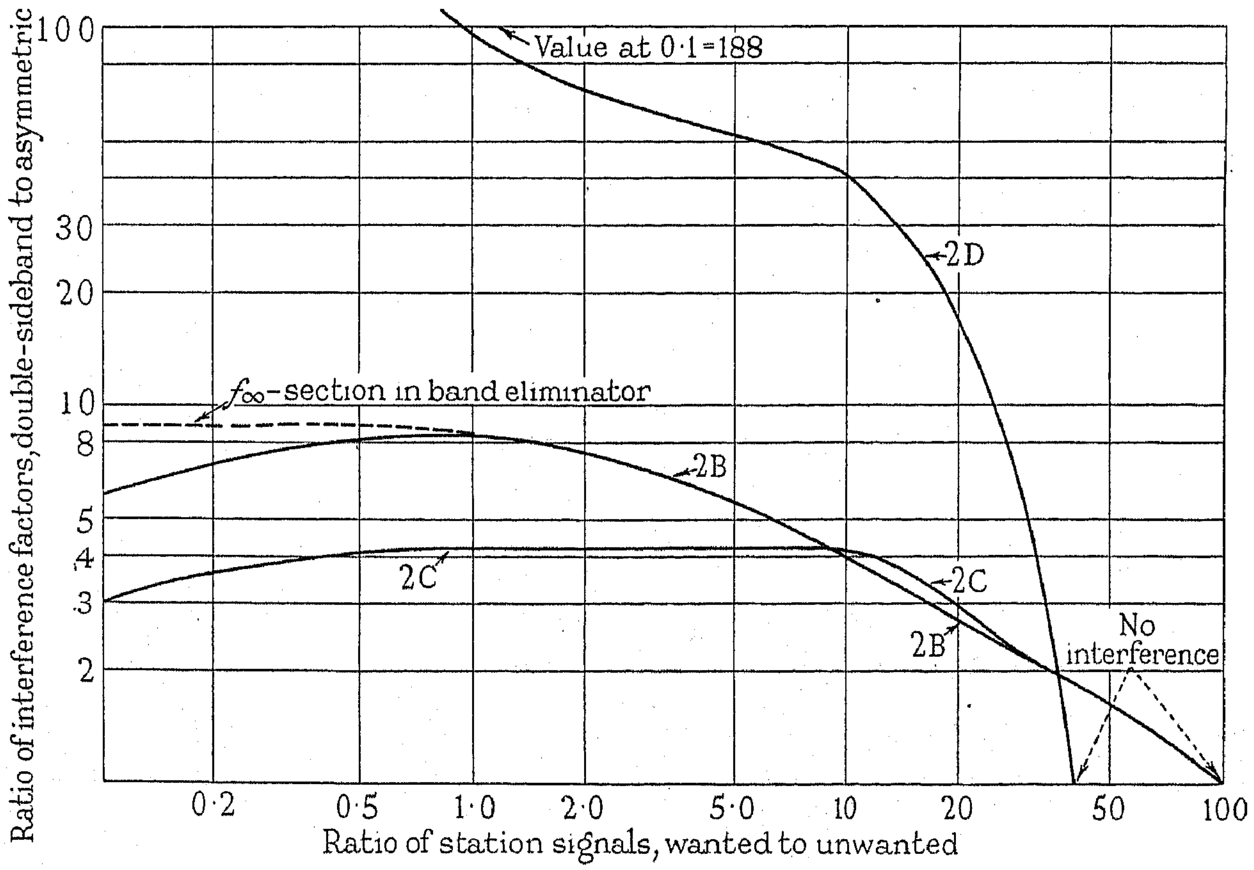

Fig. 31

width to include not only the wanted but inevitably the unwanted sideband. The asymmetric system postulates co-operation between receiver and transmitter, and this is one of the difficulties in applying it to space broadcasting.

So far we have postulated " ideal " reception, i.e. the receiver has been supposed to be responsive only to the energy components of the wanted station and completely unresponsive to any other. But evidently this is a theoretical postulate; the receiver cannot embody filters with infinitely sharp cut-offs nor filters which, with asymmetrically-disposed characteristics in relation to the received carrier, do not also introduce distortion. It is clear that the wave-envelope distortion, due to magnitude and phase asymmetry between sideband components introduced by a sideband-cutting filter, can be produced equally in the receiver and the transmitter. In other words, it is the combined characteristics of all the filters which are interposed between the circuits creating a normal modulation of the carrier and a detector which must be analysed before the degree of distortion and sideband-cutting may be assessed. It would be unnecessarily laborious to work out an example here, but in general the modifications to previous conclusions introduced by the practical receiver demand that the carrier separations should be a little greater than those given by the former analysis.

Modifications of Space-Broadcasting Receivers for Asymmetric Transmission

If asymmetric technique were applied in to-day's conditions of space broadcasting to a given transmitter, it is doubtful whether its introduction would be remarked by any of the possibly millions of listeners who are accustomed to tune in that station. The distortions produced would be negligible compared with those existing in typical receivers, and the apparent strength of signals would be the same as before the changes were made. Presumably also in certain conditions of reception the interference from a powerful neighbouring station would be the same as it was before the changes were made.

This interference could only be eliminated by narrowing the band width of reception of the receiver and disposing the carrier frequency asymmetrically within this band. This process would eliminate the interference even if the other neighbouring station created a relatively strong field at the given point of reception. The interference would, however, only be eliminated at the risk of an introduction of added harmonic distortion in the receiver. This increase of distortion would not, it is suggested, be remarked by the ordinary listener.

Turning from this robust consideration of what could undoubtedly happen to the tuning of the ordinary set by the ordinary listener who is automatically finding the best way to benefit from the new system, we may discuss what would happen to the high-quality set in the field of an asymmetric spectrum. Firstly, no high-quality set can be so described when to-day it seeks to pick up any but the local station; to eliminate sideband jamming during distant and some local-station reception necessitates the introduction of increased selectivity, which robs quality of its prime necessity, namely " top " reproduction. Secondly, if the high-quality set tuned to the local station with the carrier frequency symmetrically placed in relation to its filters is to be so good as to appreciate, without a reference standard, the deterioration of quality due to asymmetric transmission, it will be a comparatively expensive and a very rare article indeed. It is therefore obvious that the problem, if indeed it can be said to exist, of the high-quality receiver is one concerning a very small minority.

During the discussion of the author's original paper a number of persons gave it as their opinion that the problems of eliminating spectrum-overlap interference could be solved by designing the intermediate-frequency filters of the receivers so that they would accept only the carrier and one sideband, and by vastly intensifying the carrier component at the detector, transmission being meanwhile modified to a carrier-and-single-sideband system. This would be an ideal solution, of course, if some 30 million receivers could be replaced on the same day. Since this would be impossible, obsolescent receivers would, according to this suggestion, suffer distortions from the modified transmitters. These distortions would be intolerable, amounting to 10 % for 80 % modulation at every modulation frequency; not, as in asymmetric technique, to some 2 % to 3 % only at modulation frequencies between 2 500 and 3 500 cycles per sec.

Another suggestion made was that double-sideband transmission should be continued but that receivers should be adapted to cut off one sideband and intensify the received carrier so as to eliminate the distortion introduced by the receiver filters in cutting off one sideband. This, however, would be no help when the sideband chosen was interfered with by a neighbouring station. It was suggested that a solution to this difficulty might lie with a frame aerial to eliminate the neighbouring station prone to jam the sideband selected for reception, and a sideband-eliminating filter to get rid of the jamming from the other. This method attempts to turn the listener into a skilled wireless operator, a process which would be widely resisted.

It is important to note that in respect of both these suggestions authoritative opinion showed that it was perfectly feasible to design receivers which would in fact augment the carrier component in relation to the sideband component. The use of such types of receivers for high-fidelity local reception would therefore reduce even what are still by the author considered to be negligible harmonic distortion components in the asymmetric transmission to values acceptable to the most discriminating ears. In sum, the ordinary receiver seeking distant-station reception could not remark distortion in comparison with that already existing. The high-quality receiver, useful as such only when tuned to the local station, would not introduce any more distortion than that existing in transmission and proved to be negligible. The high-quality and eventually all receivers could eliminate *all* distortion if means were adopted to intensify the carrier component at the detector. This point is stressed in the Conclusion to the paper.

(8) ADAPTATION OF EXISTING TRANSMITTERS

The experiments so far conducted have been concerned with carrier frequencies of the order of 25–27 kc/sec., partly because the system was primarily devised for a carrier-frequency wire broadcasting system, partly because it is naturally far easier to shape filter characteristics to requirements for sideband attenuation at these frequencies than at " broadcasting " frequencies.

Low-Power-Modulated Transmitter

In adapting a low-power-modulation transmitter, the low-power stages, where modulation takes place, would be substituted by a low-power asymmetric-sideband modulator using a frequency of the order of 30–50 kc./sec., and this frequency would be multiplied to the value of the carrier frequency used by the station and would then be fed into the high-frequency magnifier. Naturally all the filters in this magnifier would have the carrier frequency f_0' located at their mean pass frequency f_m', and no further phase or magnitude asymmetry between sideband components would be introduced.

High-Power-Modulated Transmitter

In transmitters using the high-power system of modulation, filter networks giving the required asymmetric treatment to counterpart sideband components would have to be interposed between the high-power output from the modulated magnifiers and the input to either the aerial or, if used, further magnifying valves. Such filters would presumably have to perform their functions at frequencies coincident with or near to the carrier frequency, and, unless the proper types of network were chosen to prevent it, excessive and uncontrollable voltages might be developed across the terminals of the filter elements. Thus two problems would arise in adapting the high-power-modulated transmitters: first, how to shape the attenuation/frequency characteristics when the ratio $\Delta f/f_0$ was relatively very small; and, second, how to avoid the development of very high voltages across the filter elements. The author has not been able to give more than a superficial consideration to these problems and his existing conclusions may therefore not be verified, but it would appear that suitable networks could be devised which would approximate in their performance to the ideal and yet be practicable in use.

The great majority of transmitters in use in the world to-day employ low-power modulation, and could therefore be cheaply and efficiently modified to embody the best form of the asymmetric system, assuming that demodulation correction must take place in the transmitter.

(9) CONCLUSION

In spite of discouragement from some quarters, the public insists that the wireless set exists to give pleasure; the listener chooses this or that programme because it pleases him, rejects another because it does not. The attention that the wireless-set user gives to a programme is also partly determined by the quality of reproduction but probably more by the freedom it enjoys from interference from extraneous sounds. It is primarily programme interest which pleases, but interference is a bugbear however good the programme. If proof were needed that considerations of quality of reproduction are not paramount in the mind of the ordinary listener one would have but to listen to the sounds emanating from his set and compare them, not necessarily with the original, nor even with the quality obtainable from a high-fidelity set, but rather with the results given by a receiver costing only a little more than the most popular and which is well within the purchasing power of the majority. If proof were needed that interference is a greater bugbear to the ordinary listener than the combination tones created by harmonic distortion, an analysis of correspondence and of the efforts of national committees to cure it would furnish such proof. This is not said in disparagement of the designers and retailers of wireless sets, who rightly cater for public demand, nor need it be inferred that because public interest centres round programme content, the first cost of a set, and convenience, listeners would be indifferent to improvement in quality of reproduction.

If these observations be accepted as truly representing the situation, then it seems to be hardly justified to forgo the advantages of asymmetric-sideband broadcasting because it may introduce some small extra measure of harmonic distortion. This point is raised because in the discussion on the author's previous paper there appeared to be a considerable body of official opinion which opposed the scheme on account of its alleged failure to give an acceptable standard of quality. The quantitative analysis and the results of experiments presented in this paper should, it is hoped, prevent a repetition of this criticism.

But if the objections of that esoteric but influential few who hold that quality is of paramount importance are enough to cause the rejection of the scheme as so far promulgated, then it is hoped that they will accept it when it is put forward as a step in an evolutionary policy. No one will deny that ultimately the ideal solution is to use a carrier and single-sideband system for the broadcasting service, provided flawless quality can be obtained. Therefore the scheme proposed is one which aims to achieve this ideal gradually but which asks for some courage in facing a temporary and theoretical failure to produce a completely distortionless transmission. The first step would be to introduce some form of asymmetric transmission. The slight distortions, noticeable perhaps by a very few listeners using very rare and expensive sets and possessing highly-discriminative ears, would soon be cured by the technique of carrier intensification in the receiver. The luxury of yesterday becomes of wide application to-day; one might well see, therefore, the introduction of the asymmetric system as the stimulus towards wireless-set evolution in terms of what some call homodyne reception, namely carrier-component intensification at the detector. If a majority were so equipped then the transmitters could safely narrow their radiated spectra and increase a distortion eliminated in the receiver. In the end, and assuming a sufficient number of receivers properly designed to make use of the changed technique, single-sideband transmission might be adopted and the proper solution achieved. But apart from this suggested solution, and dealing only with to-day's problems, it would appear to the author better to risk the introduction of minor distortions than to continue with the existing and abominable interferences which are set up whenever one seeks to listen to distant stations and which are all too audible when one is listening to local stations in B and C service areas.

(10) ACKNOWLEDGMENTS

The author is indebted to the directors of David Allen and Sons, Ltd., who have allowed him to spend considerable time on the preparation of the paper, and to the British Broadcasting Corporation for providing the funds for the construction of test apparatus. Sincere thanks are due to Countess Von Zeppelin for her work, detailed in Appendix II, in evolving a most useful expression for the distortion component when both ξ and ϕ are variable. Mr. R. E. H. Carpenter has given pertinent criticism and advice during the progress of the experiments, and Mr. G. B. Ringham has greatly aided the evolution of the theories herein set out by his careful conduct of the experimental design and construction work; efficient help has been rendered in the latter connection by Chief Petty Officer G. V. Burchell.

APPENDIX I

The analysis given in this paper has purposely treated the simpler cases, the only variable being either ξ or ϕ, without deriving an expression for the distortion. This procedure is logical inasmuch as a method of attack is thereby illuminated and expressions for the simpler cases can be derived from the formula for the general case by giving one or other of the variable fixed values. It is, however, shown on pages 143–147 that the general formula has a limited although highly practical application. It is therefore proposed in this Appendix to derive expressions for the distortion in the special cases in order that the application of the general formula may be verified or modified.

$\xi < 1$, $\phi = 0$.—The resultant vector OR has a magnitude given by equation (7) as

$$\mathrm{OR} = (b^2 \sin^2 \Delta\omega t + a^2 \cos^2 \Delta\omega t + 1 + 2a \cos \Delta\omega t)^{\frac{1}{2}}$$

It is further shown that the harmonic has its maximum amplitude when $\Delta\omega t = 90°$. Thus, at this instant,

$$\mathrm{OR} = (b^2 + 1)^{\frac{1}{2}}$$

Consider Fig. 32, in which the full-line curve represents a pure sine wave. The effect of magnitude asymmetry is always to raise the value of OR beyond what it would be if no magnitude asymmetry existed. This value we call D, and it is seen that the broken-line curve, which is the distorted-wave envelope, is formed by the combination of a pure sine wave (full line) and the 2nd harmonic (shown as a chain-dotted line having an amplitude wholly above the true mean line of the rectified currents). Thus we can see that the peak amplitude of the harmonic is $D/2$, where

$$\frac{D}{2} = \frac{\mathrm{OR} - 1}{2} = \frac{(b^2 + 1)^{\frac{1}{2}} - 1}{2}$$

But if $\xi = 0$, $M_N = 1$, and $b^2 = 0{\cdot}25$; then an error no greater than 1 % (and that in exceptional cases) is introduced if we write

$$\frac{D}{2} = \frac{\frac{b^2}{2} + 1 - 1}{2} = \frac{b^2}{4}$$

The distortion factor is given by dividing the peak harmonic amplitude by the peak carrier amplitude, or

$$\delta_\xi = \frac{b^2}{4a}$$

Substituting the values of b and a in terms of M_N and ξ gives

$$\delta_\xi = \frac{M_N}{8}\,\frac{(1 - \xi)^2}{1 + \xi}$$

which is the expression given on page 145.

$\xi < 1$, ϕ *finite*.—If $D_{\phi 1}$ and $D_{\phi 2}$ are the differences when $\Delta\omega t = 0$ and $\Delta\omega t = 180°$ respectively between the amplitude of the distorted wave and the amplitude of the undistorted wave, it can be seen that the percentage distortion if not too large can be expressed as

$$\delta_\phi = \frac{D_{1\phi} + D_{2\phi}}{2[1 + M_N - (D_{1\phi} + D_{2\phi})/2]}$$

We may furthermore write, from equations (15) and (16),

$$D_{\phi 1} = (M_N^2 - 2M_N \cos \phi + 1)^{\frac{1}{2}} - (1 - M_N)$$

$$D_{\phi 2} = (1 + M_N) - (M_N^2 + 2M_N \cos \phi + 1)^{\frac{1}{2}}$$

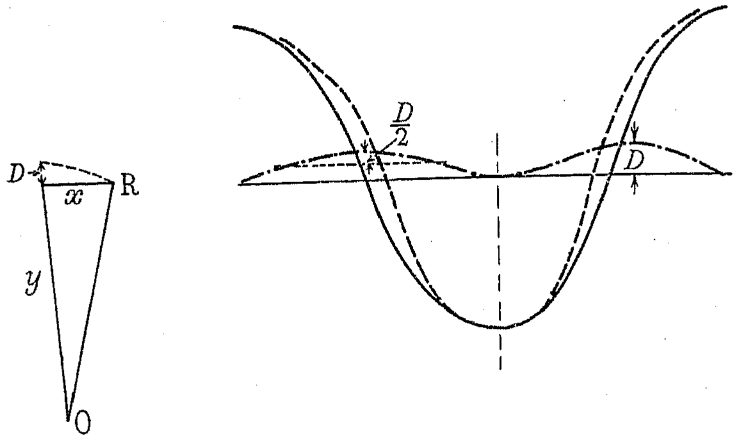

Fig. 32

This can be shown to give, for the percentage distortion,

$$\delta_\phi = \frac{(1 + M_N)^2[1 + (1 - Q)^{\frac{1}{2}}] - 2(M_N \cos^2 \phi/2 + 1)}{(1 + M_N)^2[1 - (1 - Q)^{\frac{1}{2}}] + 2(M_N \cos^2 \phi/2 + 1)} \times 100$$

where $$Q = \frac{2M_N}{(1 + M_N)^2}(1 + \cos \phi/2)$$

If ϕ is very small,

$$\delta_\phi = \frac{(1 + M_N)\,[1 + (1 - Q)^{\frac{1}{2}}] - 2}{(1 + M_N)\,[1 - (1 - Q)^{\frac{1}{2}}] + 2} \times 100$$

Points derived from this expression, which must not be used unless 2ϕ is greater than about 90°, are shown in Fig. 13B and agree well with the graphical analysis.

APPENDIX II

The case where ξ and ϕ are both variables is treated here with reference to Fig. 33, which in general is representative of the same conditions as Fig. 5 and only differs in that the attenuated sideband component in the one case is the fully represented one in the other and vice versa, and that a different instant of time is chosen in Fig. 33 from that represented in Fig. 5. Obviously the expression derived will be the same whichever Figure is used.

From Fig. 33 it is obvious that

$$\mathrm{OR} = (\mathrm{OC}^2 + \mathrm{CR}^2)^{\frac{1}{2}} = \mathrm{OC}\left(1 + \frac{\mathrm{CR}^2}{\mathrm{OC}^2}\right)^{\frac{1}{2}} \quad . \quad (30)$$

Two approximations are now made to render the algebra less cumbersome than if analysis of a complete expression for OR were attempted. These approximations are:—

(1) It is assumed that the ratio CR/OC will never exceed a value such as to make unjustifiable the approximation

$$\left(1 + \frac{\mathrm{CR}^2}{\mathrm{OC}^2}\right)^{\frac{1}{2}} \simeq \left(1 + \frac{\mathrm{CR}^2}{2\mathrm{OC}^2}\right)$$

(2) It is assumed that FC may be neglected in comparison with OF, in the denominator of the ratio $\mathrm{CR}^2/(2\mathrm{OC}^2)$.

The implications of these approximations are dealt with in the text of the paper. Using the approximations enables (30) to be written

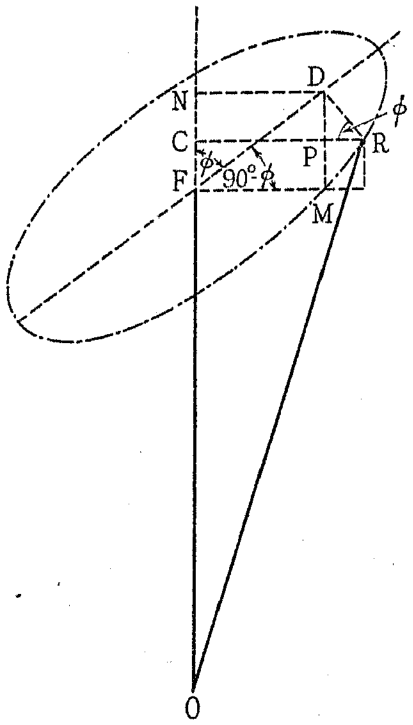

Fig. 33

- · — · Locus of R.
- - - - - Construction lines.
——— Vectors.

$$\mathrm{OR} = \mathrm{OF} + \mathrm{FC} + \frac{\mathrm{CR}^2}{2\mathrm{OF}}$$

Or, letting OF = 1,

$$\mathrm{OR} = 1 + \mathrm{FC} + \frac{\mathrm{CR}^2}{2} \quad . \quad . \quad . \quad . \quad (31)$$

It has been shown on page 142 that the magnitude of any axis of the ellipse traced by R is given by

$$(a^2 \cos^2 \Delta\omega t + b^2 \sin^2 \Delta\omega t)^{\frac{1}{2}}$$

indicating that the two components at right angles forming this resultant axis are $a \cos \Delta\omega t$ and $b \sin \Delta\omega t$.

Evidently, therefore,

$$\mathrm{DR} = b \sin \Delta\omega t \quad . \quad . \quad . \quad . \quad (32)$$

$$\mathrm{FD} = a \cos \Delta\omega t \quad . \quad . \quad . \quad . \quad (33)$$

From Fig. 33 also,

$$\mathrm{FC} = \mathrm{FN} - \mathrm{NC}$$

$$\therefore \ \mathrm{FC} = \mathrm{FD} \cos \phi - \mathrm{DR} \sin \phi \quad . \quad . \quad (34)$$

$$\mathrm{CR} = \mathrm{CP} + \mathrm{PR}$$

and, since CP = FM,

$$\therefore \ \mathrm{CR} = \mathrm{FD} \sin \phi + \mathrm{DR} \cos \phi \quad . \quad . \quad (35)$$

Substituting in (34) and (35) the values of FD and DR given in (33) and (32) respectively,

$$\mathrm{FC} = a \cos \Delta\omega t \cos \phi - b \sin \Delta\omega t \sin \phi \quad . \quad (36)$$

$$\mathrm{CR} = a \cos \Delta\omega t \sin \phi + b \sin \Delta\omega t \cos \phi \quad . \quad (37)$$

and, combining (31) with (36) and (37),

$$\mathrm{OR} = 1 + a \cos \Delta\omega t \cos \phi - b \sin \Delta\omega t \sin \phi + \tfrac{1}{2}(a \cos \Delta\omega t \sin \phi + b \sin \Delta\omega t \cos \phi)^2$$

Expanding the expression in the bracket multiplied by $\frac{1}{2}$ gives

$$\tfrac{1}{2}(a^2 \cos^2 \Delta\omega t \sin^2 \phi + 2ab \cos \Delta\omega t \sin \Delta\omega t \cos \phi \sin \phi + b^2 \sin^2 \Delta\omega t \cos^2 \phi)$$

or

$$\tfrac{1}{4}\left[a^2 \sin^2 \phi(1 + \cos 2\Delta\omega t) + 2ab \cos \phi \sin \phi \sin 2\Delta\omega t + b^2 \cos^2 \phi(1 - \cos 2\Delta\omega t)\right]$$

The value for OR can now be rewritten in terms of the sum of three different forms, namely (i) those in which $\Delta\omega t$ does not occur, (ii) those containing $\Delta\omega t$, and (iii) those containing $2\Delta\omega t$, as follows:—

$$\mathrm{OR} = \left[1 + \tfrac{1}{4}(a^2 \sin^2 \phi + b^2 \cos^2 \phi)\right] + \left[a \cos \Delta\omega t \cos \phi - b \sin \Delta\omega t \sin \phi\right] + \left[\tfrac{1}{4}(a^2 \sin^2 \phi - b^2 \cos^2 \phi) \cos 2\Delta\omega t + 2ab \sin \phi \cos \phi \sin 2\Delta\omega t\right]$$

Evidently the first expression within square brackets is a constant; the fundamental is represented by the second expression within square brackets; and the last expression within square brackets is that for the harmonic, being in terms of $2\Delta\omega t$. The amplitude of the fundamental is given by

$$(a^2 \cos^2 \phi + b^2 \sin^2 \phi)^{\frac{1}{2}} \quad . \quad . \quad . \quad (38)$$

and the amplitude of the harmonic by

$$\tfrac{1}{4}\left[(a^2 \sin^2 \phi - b^2 \cos^2 \phi)^2 + 4a^2b^2 \sin^2 \phi \cos^2 \phi\right]^{\frac{1}{2}}$$

which simplifies to

$$\tfrac{1}{4}(a^2 \sin^2 \phi + b^2 \cos^2 \phi) \quad . \quad . \quad . \quad (39)$$

The harmonic ratio is given by dividing expression (39) by expression (38), or

$$\delta = \frac{a^2 \sin^2 \phi + b^2 \cos^2 \phi}{4(a^2 \cos^2 \phi + b^2 \sin^2 \phi)^{\frac{1}{2}}}$$

Substituting the values of a and b in terms of M_N and ξ gives

$$\delta = \frac{M_N}{8}\,\frac{1 + \xi^2 - 2\xi\cos 2\phi}{(1 + \xi^2 + 2\xi\cos 2\phi)^{\frac{1}{2}}} \quad . \quad . \quad (40)$$

and (40) is the equation used, with certain reservations, in the text of the paper, to determine the distortion.

Note that from (38) we may write

$$M_A = \frac{M_N}{2}(1 + \xi^2 + 2\xi\cos 2\phi)^{\frac{1}{2}} \quad . \quad . \quad (41)$$

which is again used in the text of the paper.

DISCUSSION BEFORE THE WIRELESS SECTION, 2ND MARCH, 1938

Mr. H. Bishop: The B.B.C. has followed the author's work with great interest both from the purely technical point of view and from the point of view of its importance to broadcasting in this country and abroad. The question of the future use of asymmetric-sideband broadcasting or any other specialized type of transmission is, in fact, rather more international than national in character. Nothing is truer than the first sentence of the Summary to the paper, and it is clear that any decisions which the B.B.C may make must be bound up with what is done abroad. The wavelength situation is being debated at the five-yearly Telecommunications Conference which is at present taking place in Cairo, and we shall know at the end of the conference whether any more wavebands have been allotted for broadcasting in the face of the eager demands of practically all the other wireless services. We should indeed be foolish if we were not willing to examine any system of the kind described in the paper which might hold out some possible hope of alleviating the present wavelength congestion.

The paper refers solely to broadcasting, but it might be a good idea to examine its application to other services if there is hope of the more efficient use of wavebands for these services. So far as the present situation is concerned, the amount of interference which the broadcast listener suffers in areas of high field-strength is not serious, but, of course, in areas which it is not possible to serve so well it is a different story. On the medium wavelengths the Lucerne Plan of distribution is being adhered to fairly well by the European countries. Interference is acute in the long waveband, and in the last few years it has been particularly severe on the short waves, where no internationally agreed plan is in force.

In examining the author's system it is a question of balancing the deterioration of quality from which we now suffer, due to the attenuation of the upper frequencies of modulation to avoid interference from frequency-contiguous channels, against the deterioration which is likely to be observed with his system. In this connection I would mention the use of variable selectivity, which was raised in the discussion on the author's previous paper.

Referring to page 166, I am rather shocked by the statement that it is not the business of the transmission authorities to attempt to solve reception problems introduced by changes in transmission technique. Any change the B.B.C. proposes to introduce must take into account the receiver position, as in fact the author said in his verbal résumé of the paper. When one considers the money locked up in receivers in use in this country, it is clear that any change in transmission technique which involves changes in receivers must not be introduced without a great deal of thought and preparation.

At the Bucharest Conference of the C.C.I.R. a year or so ago Great Britain recommended that research should continue on special systems of transmission, with the intention that, when such a system had been proved to be of practical value, it might help to reduce interference and improve quality with the present frequency-separation.

Mr. H. L. Kirke: The author mentions that the receiver filter will introduce distortion, and he suggests that this distortion will be small. I should like to ask him whether he thinks that the distortion would be any smaller than the distortion introduced by a similar filter in the transmitter, and therefore why the same care has not to be taken in the receiver as in the transmitter in order that the same degree of smallness of distortion shall be achieved. He also mentioned the question of the desirability of eventually working with a single sideband and a largely suppressed carrier; I agree that that is the system to aim at. If we can get a step nearer to that ideal by the introduction of the asymmetric-sideband system or some modification of it, then I think that that may be a good thing to do, but I am not certain whether the asymmetric-sideband system by itself and for itself is good. Great difficulties lie ahead of any organization which attempts to get all countries to do one and the same thing with one and the same object; until that can be done, the asymmetric-sideband system will only be of use where two countries come to an agreement independently of the rest.

I hope that the idea of closer spacing of carriers will not be adopted. In the last few years the technique of receiver design has gone forward by leaps and bounds. It is now possible to buy a radio receiver at a not very extravagant price which will receive well up into the region of 8 000–9 000 cycles per sec., and it makes a vast difference to the enjoyment of broadcasting programmes if that band of frequencies can be received without distortion. Most of the receivers which are in use to-day have their tone controls turned to what is usually called "mellow," chiefly because of the distortion which is introduced.

I agree with Mr. Bishop that the transmission authorities must study receiver technique, whether it is a question of changing over to the single-sideband and suppressed-carrier system or, for example, of changing over from amplitude modulation to phase or frequency modulation.

I find the paper very confusing, and I feel that if the author could have referred to such quantities as M_A and M_N a little more in words it would have made it easier even for the mathematician to follow what is a very complicated subject.

Dr. E. K. Sandeman: The first point which occurred to me in looking at this paper was that it was not essential to use a band-eliminator filter for the purpose of obtaining phase compensation: other types of phase-compensating network could be used equally well.

A point arising out of one of Mr. Kirke's remarks is that asymmetric-sideband transmission demands the introduction of correction in the transmitter for the distortion introduced by the filters in the receiver. This means that it would be necessary that the same type of receiver should be used everywhere, a restriction which would impose a serious limitation on the manufacturers of receiving sets.

I feel that the difficulties of straight single-sideband working are largely resident in cost considerations rather than in the technical problems outlined in this paper, which are all sidetracked by the introduction of a large local carrier.

I also think that the practical difficulties of any system of asymmetric-sideband working are appreciable. For those reasons, and also because single-sideband working has definite advantages over asymmetric-sideband on the grounds of channel spacing, I should prefer to wait until the progress of receiving-set design and manufacture has further reduced costs to the point where the community as a whole can afford straight single-sideband working.

Mr. P. G. A. H. Voigt: When the author read his last paper before this Section he assured us that the distortion introduced by his system was negligible, and I was very hopeful of its success. This paper appears to prove that about 2 % is the maximum distortion which is to be expected if his system is treated carefully. The big difference which has occurred in the interim is shown by the fact that whereas in the previous paper the carrier wave was running halfway up the skirt of the sideband, in this paper it is on the top.

I am able to confirm to some extent the author's curve for energy in the spectrum. A long time ago, before frequency standards were available for the testing of microphones, I came to the conclusion that it was possible to derive some relevant information by striking a piano mechanically at a constant strength, and measuring the output with a microphone. If the efficiency of the piano were the same on each note a level response would be obtained, and I expected that if I found great discrepancies the trouble was likely to lie in the microphone rather than in the piano. The results of that test showed a large difference between one note and the next, but if all the results were plotted, and especially if the microphone was tried in various positions and the average value determined, a fairly level response was obtained over the middle part of the scale, and a falling-off at the extreme top. I have since been able to check that the microphone used was substantially linear to within 3 db.

Reverting to the question of distortion, the figure of 2 % is very reassuring, especially if the interference can be cut down by the amount stated. At present, high-quality reception with very flat tuning even of the London National station in certain areas in South London is completely spoilt by the fact that London National fades at times so that it is only a slight signal which interferes with Trieste. The resultant sideband splutter, especially if Trieste is over-modulated, is very bad.

For reception on the system described in the previous paper I gathered that the standard receiver would be satisfactory. It now appears that an addition will have to be made to the receiver, and that addition will in turn produce some additional distortion. Might it happen that the distortion produced in the receiving corrector would be the inverse of that in the transmitting corrector, and so compensate for it?

Mr. C. E. G. Bailey: The author's demonstration by means of his celluloid vector was a valuable illustration of frequency modulation; thus, in the case where ϕ was 90° the vectors were of equal magnitude, and the resultant hardly varied at all. It is interesting to reflect that the amount by which that resultant did vary represents the difference between a system with first-order sidebands and one with true frequency modulation. This difference would be equalized by a suitable set-up of additional vectors, and the model made to give, say, a second-order approximation of frequency modulation.

As regards the method of eliminating phase distortion, better results might have been obtained with an m-derived section, which would only have meant one extra element as compared with the band-stop filter. It might in fact be found that if the original filter were turned from a constant-k into an m-derived type the slope would be made steeper, and it would be possible to dispense with the extra cell.

I am interested in the question of the phase characteristics of the filter in the receiver. In the conventional superheterodyne receiver this filter simply consists of pairs of linked circuits. The phase therefore varies widely, and if the filter is arranged, as it frequently is in the case of sets of expanding selectivity, to have a double hump, this phase characteristic may even have a kink in the centre. When the receiver has been in use for some time and the capacitance arms of the filter have altered in value, the amplitude characteristic becomes very asymmetrical. With double-sideband transmission some of this symmetry cancels out, but it seems likely that the phase characteristic may prove worse still, and may in effect annul many of the advantages of the elaborate filter put up at the transmitter in a single-sideband system. This point should not be neglected, especially as the paper is one in which the practical difficulties of marrying a new transmission system to an existing collection of receivers have been carefully considered.

As regards the author's demonstration, I noticed a certain amount of distortion throughout his experiments.

Dr. L. E. C. Hughes (*communicated*): We know that the standard of reproduction tolerated by the public is deplorably low, and the recent advances in receiver design which make broadcast transmission more pleasant to listen to are not reflected in the mass-produced receiver. It is true that the amplitude distortion permitted in the author's system is low in comparison with that normally tolerated in radio receivers. At the same time he is retarding progress in good-quality reception by making it impossible for receivers to be used which aim at a lower degree of amplitude distortion. Such a procedure cannot be defended on technical grounds, while the complete elimination of one sideband of frequencies, no difficult task, establishes at once an implied improvement in every demodulator, and therefore is worthy of support, although it may be many years before it is practicable to re-allocate international frequencies to permit the operation of more channels. It is nearly impossible to contemplate the elimination of the carrier, in spite of the power savings at the transmitter. Engineers cannot defend giving the public what it is alleged to want, because the public has no possible means of judging what it might want if it knew what was available. The task of enlightening the public to an appreciation of good-quality reproduction, and what it is worth, might be considered a legitimate activity of the new Engineering Public Relations Committee, of which The Institution is a supporter. If the public could be persuaded to take broadcast programmes more seriously, broadcasting, as a system, would become more valuable and so justify its great expense. One of the major factors is the acoustic fact that good-quality reproduction, involving crisp speech and clear-cut acceleration of notes, means a clean bass and good high-note response without a sharp cut-off, and is far more attention-compelling than the " mellow " tone which the public has been trained to accept, by the simple process of not offering it anything else. The great majority of the public has not been trained to listen to original music, and does not attend orchestral concerts or appreciate high-grade violin playing at first-hand. I have the opportunity of listening to between 50 and 100 new broadcast receivers and loud-speakers every year under carefully standardized conditions, and I find ample evidence that the general quality of the mass-produced broadcast receiver is deteriorating. The trade is not passing technical benefits to the public, and from this point of view any advocacy of further degradation is retrograde, whatever technical merit the author's scheme otherwise possesses.

Mr. W. T. Sanderson (*communicated*): The paper has justified, by analysis, the faith the author expressed some 2 years ago, in the practicability and advantages of his system of asymmetric-sideband transmission. Whether it is desirable at this stage to press for its adoption depends on one's opinion of the future of broadcasting as a whole.

There is in many minds a belief that all high-quality broadcast reception will eventually be achieved by the use of wires and that wireless broadcasting will continue principally to serve large, thinly-populated areas. For this purpose and for all the commercial services occupying the other available channels, a distortion of a few per cent is unimportant and the asymmetric system offers most valuable advantages.

Even if broadcasting were to continue on the present lines, I should hesitate to support the immediate adoption of the asymmetric system only because of the fear that, as soon as the change had become effective, station separation would be reduced to 5 kc. and all chance of getting good reproduction wiped out for ever.

It is claimed that the public do not want the reproduction of top notes because if a tone control is provided the user always sets it at " mellow." Surely this is a false deduction. Much as one may dislike " mellow " reproduction, it is more pleasant than the harmonic distortions, resonances, and sideband splash, which contribute so largely to top-note response to-day. Further technical advances, particularly in loud-speaker design, should remove present difficulties, when the demand for a wider frequency spectrum would probably arise. Asymmetric-sideband transmission would provide a means of satisfying the requirements if a 9-kc. carrier separation still existed. A large proportion of music is written to stimulate the hearer, and the higher frequencies produce such stimulation. The average listener may, however, want to be lulled, and may use the tone control to this end.

Mr. P. P. Eckersley (*in reply*): The asymmetric-sideband system was designed to eliminate sideband-splash interference, not, as would appear from the discussion, solely to introduce harmonic distortion.

The decision as to whether the system merits adoption would be more easily taken if some organization would spend the few hundreds of pounds necessary to make a trial.

In the absence of large-scale practical tests and continuing discussion instead of much-needed action the merits are debatable, as Mr. Bishop states, in terms of a balance of factors. The discussion has revealed a quite understandable reluctance on the part of responsible people to introduce even more distortion into broadcasting systems, but has not shown any desire to consider the merits of asymmetric technique in regard to its ability to eliminate interference. As a matter of principle it is right to attempt, so far as is practically possible, to eliminate harmonic distortion. As a matter of principle it is equally right to attempt, so far as is practically possible, to eliminate interference. In principle, therefore, these two factors cancel and it is the harder task, once the safe shelter of principle is abandoned, to decide the question quantitatively rather than qualitatively.

No one has challenged the quantitative conclusions set out in the paper and they can therefore be re-stated with more confidence since they have passed unscathed the critical examination of well-informed technicians.

(1) The harmonic distortion introduced in transmission by asymmetric technique amounts to between 2 % and 3 %. This maximum figure of distortion occurs only when modulation is of the order of 80 %, namely it occurs during about 10 % of the time of transmission. The distortion of this value occurs at a modulated frequency of 3 000 cycles per sec. (roughly) and falls to negligible values at 2 000 and 4 000 cycles per sec. and is thus concentrated in a narrow band of frequencies.

(2) For the introduction of the negligible amount of distortion in transmission defined above, sideband splash may be reduced 10-fold, but receivers, when tuned asymmetrically to do this, may introduce more distortion than the 2 %–3 % existing in transmission.

As a reassurance of (1) above it has been proved that even highly trained ears have found considerable difficulty in distinguishing between orthodox and asymmetric technique even in terms of a standard the public could never expect unless they were allowed a service of wire broadcasting.

Thus, in the first place, high-quality reception need not be disturbed by the introduction of the asymmetric-sideband system, because high-quality reception can in any case only take place when the field from the wanted station is very strong and when therefore sideband splash is not heard and when therefore the receiver may be tuned symmetrically. The symmetrically tuned receiver does not introduce further distortion due to asymmetrical treatment of sidebands and that is why, in answer to Mr. Kirke, it is necessary to take so much care in designing the transmitter filters, since they alone determine distortion in conditions for high-fidelity reception.

But the point made by several speakers is that the receiver must be tuned asymmetrically if it is to benefit from the introduction of asymmetric transmission in eliminating sideband-splash interference, and it must, in being tuned asymmetrically, introduce more distortion than is present in transmission.

The question is then: " Is it worth while introducing more distortion for the sake of eliminating interference? " This question is resolved by knowing how much extra distortion will be introduced, or more exactly by knowing what proportion of the total distortion existing in typical receivers will be added by asymmetric tuning and asymmetric transmission. In the latter question we are involved with Dr. Hughes's point that, bad as receivers are, we should encourage improvement rather than, as he implies, discourage it by introducing more distortion.

Taking these points in order of their statement, I think that, provided it can be proved that the extra distortion, in spite of asymmetric tuning where sideband splash is otherwise inevitable, is negligible, then the asymmetric system should be used.

Mr. Voigt has brought out a point which I think is highly relevant in asking if in fact one must always refer to " added distortions." In any broadcasting system we must presuppose several sources of harmonic distortion. Is it conceivable that the phases of all these distortions will be coincident? Surely not. Indeed, experiments prove that the amplitudes of harmonics seldom add arithmetically. I have attempted at different times to prove experimentally the findings of my analysis, and seldom have I been able on a first experiment to get as much harmonic as would be indicated by theory to exist, given certain measured degrees of asymmetry between counterpart sideband components. Only after a rigid examination of the rest of the system to eliminate all sources of harmonic distortion have the quantitative results agreed with theory, and, as said above, the sum of a number of harmonics is a vector not an arithmetic addition.

Mr. Bailey observes, and I see no reason to deny his contention, that phase asymmetry in typical receiver filters may introduce vastly more distortion than is spoken of as a part of asymmetric-sideband technique. But I have lately made experiments which prove that the addition, in the chain of filters between modulation output and detector input, of two filters consisting each of two tuned circuits coupled through the mutual inductance between their inductance elements does little to change the M_A versus Δf characteristic, showing that the extra phase distortion is very small.

The proper adjustment for the receiver filter is that it shall be substantially flat over a range of frequencies $f_0 - \Delta f_\infty$ to $f_0 + \Delta f'$, where $\Delta f'$ is the highest modulation frequency fully represented.

If this adjustment is made there is a moderate degree of phase-characteristic symmetry, and that is probably why the extra distortion introduced by a receiver filter so properly adjusted is slight.

Dr. Sandeman and, to a degree, Mr. Kirke miss this point when the former argues that every receiver will not only introduce extra harmonic distortion but extra demodulation as well, making it necessary to apply demodulation correction to every receiver or to use the same type of receiver " everywhere." If the receiver filter is flat over the range $f_0 - \Delta f_\infty$ and only introduces an attenuation, which tends to eliminate a sideband component already substantially removed in the transmitter filter, at frequencies lower than $f_0 - \Delta f_\infty$, no extra correction is needed in these conditions since the full 6-db. correction required has already been supplied in the transmitter.

My own feeling is, that, seeing that there is a good deal of distortion present in transmission in any case, believing that distortions do not necessarily add but often cancel, and knowing that there is a large amount of distortion existing in receivers in any case, the extra distortions, due to asymmetric tuning, would pass unnoticed, whereas the elimination of interference would not; the former, even if present in a small degree, is fully compensated for by the absence of the latter in large degree.

We come now, however, to a point implicit in a statement made by Mr. Bishop and clearly set out by Dr. Hughes. This is the question as to whether we should not leave things alone and attempt rather to encourage the production of better receivers than to recognize distortion as inevitable and to introduce more than is absolutely necessary in transmission. The shock administered to Mr. Bishop by my statement " that it is hardly the business of transmission authorities to solve reception problems introduced by changes in transmission technique " rebounded upon its author when I realized that this was a quotation outside its full context and Mr. Bishop failed to state that the sentence he quoted was completed by saying " provided they can be satisfied that it is not impossible or impracticable to design circuits to make full use of the changes." I would remind Mr. Bishop that he was as impatient as I was when permission to proceed with the Regional Scheme was delayed because it was alleged that it was impossible for a typical receiver to select between two transmissions of equal strength having an average of 100 kc. carrier-difference frequency. My statement, when taken in its full context, seems justifiable unless all of what used to be called " private enterprise " is to be managed by the State.

In spite of Mr. Kirke's revelation that for a moderate price one can buy receivers " which will receive well up into the region 8 000–9 000 cycles per sec." (i.e. into the region of sideband splash), I am inclined to agree with Dr. Hughes " that the general quality of the mass-produced receiver is deteriorating." Mr. Kirke's receiver would seem to be chiefly useful for local station reception, but I would add that it has to be a very good receiver even for that purpose if, as Mr. Sanderson so rightly observes, its top reproduction is not to be more irritating than pleasing. The average receiver wisely cuts off top to avoid sideband splash and, on local station reception, that horrible hard grating noise that is so prone to accompany a receiver with a " good " frequency characteristic.

Turning, however, to Dr. Hughes's contribution, one is forced to be drawn into seemingly irrelevant discussions of economics and sociology if one is to prove that, admirable as are his intentions to elevate public taste, the means to do so in the present state of affairs are non-existent. While it is no doubt deplorable that the standard of reproduction tolerated by the public is very low, it is equally sad that the cost of good reproduction, both in money and in bulk, is very high.

I agree that it is the duty of engineers to attempt to " pay the public the compliment of offering them something slightly better than the public thinks it wants," but it is no good doing this unless, as the salesman says, " the price is right." Reforming zeal seems often more pronounced as the problem to be solved is more removed from reality. The way to give the public good reproduction is first to create a public which demands it, and secondly to give them the money to pay for it. Deploring the fact that private enterprise produces a cheap article which makes a nasty noise but which is the only article the average public can buy, gets us no farther. Is it, moreover, realized that besides being costly the apparatus of good quality is extremely bulky? People tend to live in smaller and smaller rooms, while a wireless set of the high-fidelity type is relatively massive. The " bijou set " was sold in such vast numbers mainly because it was so " bijou," cost very little, and gave an adequate result. It supplied a public demand. Before, therefore, the engineer can do his duty he must face the facts, which are (1) that the public wants to be amused, (2) the public either does not find broadcasting sufficiently amusing to justify a large expenditure, or (3) has not much money to spend on it, or (4) even if an individual has a great deal of money he nevertheless does not want to lumber up his rooms with great boxes emitting even the most crisp top.

In sum, assuming the continuance of the present economic and political regime and assuming the continuance of wireless rather than wire broadcasting, distortion in the majority of receiving sets is inevitable. Desirable as it would be to reduce it, I cannot see how it can be done without State control, regimentation, and subsidy. Assuming, therefore, that distortion, in large amounts, is inevitable, I argue that the extra distortion, which may well be a less distortion, due to asymmetric transmission and reception will not be noticed. Which then is worse, a distortion which would not be noticed or an interference that is? Is quality alone defined in terms of harmonic distortion? Does not the ridding of interference constitute a gain in quality? Is not the possibility of extending the frequency range an increase of service?

Nor is this all. The asymmetric principle, according to my proposals, is the first step in an evolution towards the commonly desired end of carrier and single-sideband working. I cannot accept the " reasons " which led Dr. Sandeman to deny the relevance of my proposals to treat the asymmetric system as a step in an evolutionary development. Dr. Sandeman would prefer to wait until the progress of receiving-set design . . . " has further reduced costs to the point where the community as a whole can afford straight single-sideband working." Dr. Sandeman would be more convincing if he even indicated what this evolution should be, how receiver manufacturers, interested only in making money, are going to tackle the problem of producing receivers for " straight " single-sideband working, and what incentive there is to produce single-sideband receivers for double-sideband transmission.

I greatly appreciate Mr. Voigt's contribution to the discussion, which reinforces the ideas I have endeavoured to set out. The experiments he made on the sound spectrum analysis are interesting in that they confirm the general usefulness of the curve of Fig. 14.

Turning, at the conclusion of this reply, to the more technical points, I agree with Dr. Sandeman that it would be possible to use a multi-section filter in conjunction with phase-compensation networks, but would it be possible to produce an apparatus having the same performance as the one described in my paper for the same cost? It would seem to me that a merit of my system lies in its simplicity and therefore small cost. I agree with Mr. Bailey that *m*-derived sections could be used, particularly in the band-elimination filter, in order to continue the full attenuation over the outer parts of the cut-away sideband, but it is perhaps proved by the spectrum-analysis diagrams that the present performance could hardly be bettered in the generality of cases without introducing either more distortion or more spectrum overlap.

I am grateful to Mr. Bailey for showing how my model, designed to show the vector analysis of the asymmetrical treatment of sideband components, could be adapted to illustrate frequency modulation—the point had escaped me.

Bad distortion was, as Mr. Bailey observes, present during the demonstration given after the meeting, a fault (later discovered) having developed in the receiver common to both transmissions. I am sorry that because I adopted the commonly-employed convention of setting out the definition of the symbols in an introductory list, rather than defining them each time they were mentioned, Mr. Kirke found more difficulty in understanding the paper than is implicit in the mastery of the subject. Mr. Kirke should sympathize with my difficulties, however, when he appreciates that several critics said the paper was far too long while he, on the other hand, would have it longer.

In conclusion it is worth repeating that the asymmetric system was developed primarily for use in connection with carrier-current wire broadcasting, wherein it has proved both practicable and useful.

As Mr. Sanderson so relevantly observes, wire broadcasting must obviously one day replace much of wireless broadcasting, and maybe it would be better to concentrate on developing this system which, for a cheap price, would give good-quality reproduction to the public, than spend more time and money on " bolstering an inefficiency " by applying asymmetric-sideband technique to space broadcasting.

HIGH-POWER VALVES: CONSTRUCTION, TESTING, AND OPERATION*

By J. BELL, B.Sc., J. W. DAVIES, and B. S. GOSSLING, M.A.

[Communication from the Research and Engineering Staffs of the M.O. Valve Co., Ltd., at the Works, Hammersmith, and (G.E.C.) Research Laboratories, Wembley.]

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SUMMARY

In the Introduction the authors discuss how the development of high-power valves is influenced by the standard of reliability demanded by such services as broadcasting.

The principles governing the constructional design of each main part of the valve are outlined; differential expansion frequently appears as a controlling factor.

The glass-work is of lead-potash-soda glass, the anodes of thick copper tubing; for glass-to-metal seals nickel-iron alloy is preferred, but copper is sometimes used.

Insulating members have been eliminated from the active part of the valve. Grids are not cross-braced, grid seals include a large ring section let into the bulb, and there is an 8-lead multiple seal for pentodes.

Cathode seals for 1 000 amperes are described; these support the whole cathode system, the evolution of which into a free-hanging multiple construction is outlined. Heat transfer between anode and cooling liquid is discussed, and also forced-air cooling.

Modern evacuation technique proves to be governed by two effects; the lower readings of grid current in the "gas test" are shown to be due to photo-electric electron emission caused by X-rays from the anode and not directly to gas, and the clean-up capability of the valve is found to be very large.

The method is described for determining the operating filament voltage for a standard emission by extrapolation to full emission from a reduced-emission test. Examination of the statistics of emission-test data shows that former variations are to be ascribed to variable thermal emissivity and not to variable dimensions.

Methods are given for extrapolating low-power space-current readings into the operating region, with allowance for division between anode and grid, and the "tail" of the anode-current characteristic is discussed. The control of the secondary-emission component of the grid current and the effect of the magnetic field of the filament are described.

An account is given of later experiments and of recent experience with flash-arc breakdowns (Rocky Point effect).

Recommendations are made for switching-on filaments and anode potential in operation, and for purity of cooling water.

Curves are shown of the distribution of evaporation wastage in various types of filament. Typical valve-life data from normal operation are illustrated by survivor curves for constant voltage and for constant emission during life, and a Table is given showing the recent performance of the largest valves at three stations.

A table of ratings for the various types of valve covered by the paper is also included.

INTRODUCTION

The development of high-power valves is a relatively slow process, so that it is only at intervals of several years that advances can be recorded in any comprehensive fashion. It is actually some 10 years since valves of the particular sort here considered—those made by the M.O. Valve Co.—were described to this Institution by W. J. Picken.† This slowness seems in retrospect to have been imposed by the circumstances in which the majority of such valves are used.

For a better understanding of these circumstances the following instances may be cited.

First of all there is the very high standard of reliability demanded by those services of which broadcasting is the type. In these there is no answering-back or possibility of repetition, and a brief interruption may affect irrevocably a million listeners instead of merely a few. The standard is comparable with that of power-supply plant, for although there are admittedly frequent opportunities for servicing in the intervals of programmes, there is on the other hand no automatic alternative source of supply, even temporarily. It is not the fact that a valve has failed and has to be replaced that is important; it is that an interruption to the service has occurred. It is a question of operation, not of economics. As a result of this requirement, improvements which may affect performance in operation are generally introduced one at a time, leaving well alone for the time being in other directions. Certain kinds of improvement, e.g. in mechanical construction or characteristics, have latterly been accelerated by preliminary studies, using demountable models.

Then there is the long life expected of the valves. This is not the same thing as reliability, for reliability depends chiefly on the absence of premature failures, whereas a long average life has virtues of its own. Apart from the cost of the valves themselves, there would be additional calls on the time of the station staff, and need for more spares at remote stations, if replacements were more frequent. This long life fixes the time unit of the development process at 3 or 4 years, since that is the period generally required to work through the whole of even the first batch of a new type of valve, and so to obtain a first general idea of its overall performance, although the proportion of premature failures, if any, will already have been known after the first few months. Extended life tests of valves of the highest order of power are not feasible, although from time to time manufacturers have recorded their thanks to various authorities for facilities for preliminary operation, which have been of great value in eliminating causes of premature failure. Fortunately, premature failures have been less common in recent types of valve.

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* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 176.

† See Reference (1).

Further, none of these large valves are made in very large numbers, so that the picture of overall performance must be statistically crude if there are many points of difference between different types.

Now these operating conditions and requirements which have been enumerated are much the same for all the many types of valve which come within the scope of this paper. In consequence, there are various threads of common practice running through all the several designs, and it is our purpose rather to follow these threads, giving particular instances in illustration, than to go into detailed description of the few types for which space would permit such treatment.*

To economize space by avoiding repetition, the reader is referred from time to time to a general survey of the development of large radio transmitting valves by Le Rossignol and Hall.†

CONSTRUCTIONAL PRINCIPLES

(a) General

The use made of engineering principles in the construction of small valves is largely concealed because these principles are mainly applied to the machines and tools by means of which the valves are made, and they therefore appear only indirectly in the finished valve. A high-power valve, on the other hand, shows itself more directly as an engineering construction. The principles followed are, however, distinctly specialized and must be classed as an extreme case of what may be called high-temperature engineering.

This is for the reason that in many of the parts of the valves the ranges of temperature experienced between idle and operating conditions are wide, and in some cases very wide indeed. Also, under the temporarily more severe conditions of manufacture, they are in general wider still; in fact, they include the widest at present known in engineering.

These temperature ranges imply corresponding expansions, and our experience with these valves has shown that the one overriding rule which must never be violated, if trouble is not to follow, is that the effects of differential expansion must on no account be ignored.

The valve is, of course, also an evacuated high-voltage device, but the requirements arising in these directions have been found to be of relatively minor importance as regards the constructional design.

Throughout the account now to be given, the demands of differential expansion can be seen in the background of the design, whatever part of the valve may be under immediate consideration.

Fig. 1 (see Plate 1, facing page 184) is a radiograph showing the internal and external construction of a high-power valve, type C.A.T.14; most of the features now to be discussed are indicated in the Figure.

(b) Glass

The glass envelope or bulb is a convenient starting point for this outline of the principles of construction, because in one way or another all the other parts are mounted upon it.

Throughout the development of high-power valves we have continued to use a relatively soft glass. This has served us so well that there has been little incentive to change it, and no change at all has in fact been made. The glass contains a fairly high proportion (30 %) of lead, and in consequence it is elastic and withstands " thermal shock," i.e. sudden local expansion, relatively well; it is also viscous over a rather wide range of temperature, does not de-vitrify quickly through chemical instability, and can be fully annealed at a convenient temperature. Further, joints between glass and glass are in no danger of minute pin-hole leaks. These properties form a good basis for somewhat ambitious glass-working operations, such as main glass-to-glass seals 5 in. in mean diameter.

Troubles which might arise from electrolysis of hot glass parts have been avoided by substituting potassium for half of the sodium and by adding lead. The proportion of lead is sufficient to absorb X-rays to a useful extent, particularly where, as in these valves, such rays strike glass 2 or 3 mm. thick at glancing incidence. The radiograph, taken at 150 kV peak voltage, illustrates this point.

The presence of lead and the low sodium content reduce the high-frequency power factor to a low value, but no use is actually made of this virtue.

The lower softening temperature and higher expansion of this glass as compared with harder glasses of the boro-silicate type are fortunately compensated for by the fact that it absorbs much less low-temperature thermal radiation than those glasses, and is therefore subjected to smaller ranges of temperature during operation of the valves.

The thermal expansion, 9·1 parts per million per deg. C., is probably too great for constructions involving stout metallic members embedded in thick pieces of glass. However, when the very different elastic properties of metals and glasses in general are taken into consideration, such constructions, whatever appearance of robustness they may present, inspire anxiety rather than confidence, and an anxiety which is renewed with every increase of size.

(c) Anodes

The material of the anodes is copper of good ordinary quality with a low content of volatile impurities such as arsenic. In the larger valves the anodes are 5 in. in diameter and $\frac{3}{16}$ in. thick, or 8 in. by $\frac{5}{16}$ in., these latter being milled to a depth of $\frac{1}{8}$ in. in longitudinal grooves $\frac{1}{8}$ in. wide so as to double the surface presented to the cooling fluid. To prevent fine leaks exposure of the " end-grain " of the drawn or rolled metal is avoided. Leaks through the copper are very rare.

For soldering the end plates and alloy rings, high-quality silver solder free from volatile impurities is used. Actually, there was more trouble in times long past in making sound vacuum-tight joints between copper and alloy than between alloy and glass; it was found essential to distribute the strains due to differential thermal contraction fairly between the two materials.

It is sometimes found worth while to blacken parts of the inside of anodes with graphite, in order to reduce the furnace effect of the intense cathode radiation and the proportion of such radiation reaching the glass envelope.

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* A table showing the principal ratings for the various types of valve mentioned in the paper and for the other high-power types is given at the end of the paper.

† See Reference (2).

These large anodes expand to a notable extent when heated. For a 21-in. length the expansion at the " baking temperature " of 400° C. is 0·14 in., and at the full-load operating temperature of about 130° C. 0·034 in. This latter expansion can be met by using rubber rings of circular section between anode and jacket, but *not* by flat rubber gaskets.

The cooling of anodes will be considered later in this Section.

An internal guard-ring overlapping the seal is inserted into the throat of the anode for the purposes of protecting the seal and the neighbouring part of the glass from cathode radiation, metallic deposits, strong electric fields, and damage from discharges, or electrolysis consequent on direct electron bombardment.

(d) Glass-to-metal seals

The seal between anode and bulb is the most striking in point of size, but, as we shall see, the same principles apply to the other seals.

Wherever possible a ring of nickel-iron alloy whose expansion is closely matched to that of the glass is interposed between glass and copper. This ring is usually copper-plated, and treated with borax to improve adhesion to the glass. The correctness of the composition of batches of alloy is carefully watched, and all batches are checked for expansion. Given this control, the actual sealing edge need not be thinner than can readily be obtained by simple turning—and the alloy machines quite well—and it can be left with ample strength to resist the atmospheric thrust which for a seal of $5\frac{1}{4}$ in. diameter amounts to 20 lb. per inch of circumference, i.e. more than 300 lb. in all.

The reliability of these seals is unquestionable; failures, whether during manufacture of the valves or afterwards, are extremely rare—a result well worth the few additional operations they require.

Since the glass is applied to the outside only, protection from indentation of the alloy ring is provided in the finished valve. Otherwise there is only one likely cause of fracture. This should be mentioned because it might not be expected. A clean crack round the body of the glass clear of the seal itself can result from a sharp blow on the anode or its supports due, for instance, to sudden contact with a hard object. A similar blow on the glass does not produce this effect.

In cases where the seal is in a strong high-frequency magnetic field which can induce toroidal eddy-currents in it the bare nickel-iron alloy is unsuitable; it overheats to a remarkable degree. To avoid this, either the nickel-iron is copper-plated thickly enough to survive oxidation when the seal is made, or else a direct copper-to-glass seal is substituted. The latter is successfully used up to $5\frac{1}{4}$ in. diameter; it requires close control of the sharpness of the edge combined with sufficient thickness to withstand the atmospheric thrust even when the copper is fully annealed; the glass is usually applied to the inside only.

After experience of both types the ring of intermediate alloy is still preferred wherever there is free choice, except perhaps in small sizes less than $\frac{3}{4}$ in. diameter.

(e) Insulation

The glass envelope itself is the main insulator, but its insulating properties call for little comment. On the outside it is warm and dry, and high-voltage breakdown would occur first in the heated air. The only danger point for electrolytic conduction would be between the filament seals, particularly if a.c. heating is used, but there the temperature is kept low for the sake of the seals themselves.

However, those parts which are subject to high-frequency electric fields can be destructively overheated if their surfaces on the vacuum side are coated, even discontinuously, with a conducting film whose resistance is of the same order as the capacitative impedance of the space round about the film. Since the heating results from the diversion to the film of the displacement current in the space, this effect is more troublesome in short-wave valves. The formation of such films by volatilization or sputtering of internal metal surfaces during exhaust has therefore to be avoided. High-frequency dielectric losses in the glass itself are completely negligible in comparison with those in such films. The internal guard-ring extension of the anode prevents the formation of films in the stray field close to the anode seal.

Of other insulators besides the envelope there are hardly any. Their presence within the active portion of the valve is nowadays in no circumstances permitted. The combination of some or all of the adverse agencies of high temperature, high voltage, and high frequency, together with film formation, is altogether too hard on them for good reliability to be expected. There is a further less obvious probability which our experience has sufficiently demonstrated in those cases where we have tried to use steadying insulators in such places. This is, that any breakdown of films will result in much more serious breakdown of the vacuum insulation through piloting of flash-arcs—assisted Rocky Point effect.*

The grid-cathode insulators, which, if used, are of silica and similar in form to those of ordinary sparking plugs, are therefore withdrawn to a position as far as possible from the active regions, and the cathode system is insulated solely at the seals.

(f) Grids

The main constructional problem in grid-building is differential expansion. Owing to differences both in the thermal capacity of different parts, and in their rates of heating by radiation and bombardment, this sets up strains leading to progressive distortion, unless the structure is so designed as to avoid such strains. For this reason, diagonally-braced grids were early given up and an adequate, though reduced, rigidity was provided by thickening the longitudinal members.

If the μ of the valve is to be low, the grid consists simply of parallel rods; for higher values of μ a spiral of suitable pitch and thickness is added, secured by lacing spirals or, more recently, by welding. Grids in high-power valves are at present made wholly of molybdenum. The possibility of substituting tungsten for the purpose of reducing primary thermionic emission from portions overheated by radiation or bombardment has so far been held in reserve.

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* GOSSLING, see Reference (8).

(g) Grid seals

In the largest short-wave valves, e.g. C.A.T.17, the grid seals have to have a large high-frequency current rating of 100 amperes or so in normal operation, and also a considerable margin to cover mistuning of the load and local concentration of high-frequency current. The construction of the seal has therefore to be so substantial that it may as well be used for insulation and support; in these types it therefore consists of a metal ring separating two parts of the bulb with a glass-to-metal seal on either side. These are the large $5\frac{1}{4}$-in. copper-to-glass seals already mentioned; copper-plated nickel-iron is also used.

When separate seals are used, the difficulty of so disposing the external conductors as to ensure equal distribution of high-frequency current restricts to two the number of seals which can usefully be considered. Given equal distribution, the inductance of the grid leads is of course halved. These separate seals are made like miniature anode seals, and direct copper-to-glass joints are now usually used, since their special troubles are least at the sizes required, whereas nickel-iron seals have to be copper-plated, and can be troublesome if too small.

Even in long-wave valves, grid seals of good high-frequency current rating have to be provided as a safeguard against emergencies arising from parasitic short-wave currents.

For the highest-power valves of the multiple-electrode class, such as pentodes, it has been found possible to mount a ring-cluster of separate seals in the foot-tube. A pair of seals at opposite ends of a diameter is provided for each internal electrode, so as to halve the inductance, which is further reduced by keeping all seals short. For the assembly of these clusters special glass-blowing appliances and technique have been developed.

Another possible form of grid seal, not yet extended to the largest valves, is the multiple-wire seal used in some of the smaller short-wave and ultra-short-wave types. This consists of as many wires as may be desired set in a ring, and passing either through the main glass-to-glass seal between bulb and foot-tube, or through a circular " pinch " on the foot-tube itself.

This sort of seal has been found to give a very good combination of high current-rating, and low grid-lead inductance, in consequence of which the types in which it is used have given lower wavelength limits, and better performance in that neighbourhood than were anticipated.

(h) Cathode seals and mountings

The cathode seals in large valves have nowadays to carry heating currents of up to 700 amperes in operation, and considerably more during exhaust. In addition, the elimination of all internal insulators between the cathode and other electrodes requires that these seals should be strong enough to carry the whole weight of the cathode system, that is up to $1\frac{1}{4}$ lb., with the centre of gravity 24 in. from the seal. A satisfactory solution of this double problem has been found in the use of a seal which is a miniature replica of the well-tried anode-to-bulb seal. The smallest of these used is $\frac{1}{2}$ in. in diameter for 250 amperes. There has been no difficulty with progressive increase in size; the largest so far have a diameter of $1\frac{1}{4}$ in. for 1 000 amperes or more.

For sealing to the two tubular extensions provided on the main foot-tube, the two seals are mounted on a common jig of nickel-iron, which has a spring grip to prevent temporary strain in the seal when the glass is setting. The stout copper extensions of the seals on the vacuum side have accurately-machined faces, to which the internal leads supporting the cathode system can be bolted in good alignment without further adjustment. It should be mentioned here that electrical contact between clamped metal surfaces is much better in vacuum than in air. In general it can be said that these seals behave as well as they do, because the design implies a fair division of responsibility between the mechanic and the glass-blower.

A further important but not so obvious function of the cathode seals and external leads is to remove the heat conducted back from the incandescent cathode, and also that generated by ohmic heating of the internal leads, which are designed as a compromise between these two effects. The total power to be dissipated amounts to 150 watts in each 700-ampere seal. There are two alternative methods in use, the first used was liquid cooling of the hollow external leads down to the point where they are screwed into the seal. The screw, incidentally, is greased to preserve the contact. The leads have to be provided with flexible centre sections of corrugated metal tubing in order to reduce strains on the seal, and in some of the earlier large valves water-leakage troubles have developed, usually after some 5 000 to 9 000 hours, through corrosion, possibly electrolytic, of these flexible " bellows." With the silver bellows more recently used this is unlikely. The other alternative, which is in several ways more convenient, particularly when the cathode is not at earth potential, is to substitute cooling by forced air-blast. With leads having fins machined on their inner ends, and air supply at 4 in. level of water pressure, this method seems quite satisfactory.

(j) Cathodes

We now come to the innermost component of the valves, the cathode system. The final form of this, simple though it is, is yet the product of a long evolution, in course of which the essential principles were successively appreciated.

To obtain an emission current of 165 amperes, which is the largest so far demanded, with a working life of the order of 10 000 hours, a total area of some 50 sq. in. (300 cm^2) of incandescent tungsten has to be provided. This takes the form of an assemblage of filaments of diameter 1·15 mm., requiring some 22 kW to heat them, and all mounted with a constant separation of 9 mm. from the adjacent wires of the grid.

The principal mechanical factors governing the design of these multiple-cathode systems are the unavoidable facts that in coming up to incandescence the longitudinal expansion of the filaments themselves is 1·2 %, i.e. 4 mm. on a length of 350 mm., and that the lateral magnetic forces between the individual currents of 60 amperes or so tending to distort the system are by no means negligible, and are considerably greater than the electrostatic forces acting on the system when the emission is cut off or otherwise not space-charge limited. For dealing with the expansion, sliding guides are best avoided, especially if the sliding surfaces are at anything near red heat, because of the strong tendency of metallic surfaces to weld together under such conditions. If one of the surfaces is vitreous or ceramic, this tendency is probably less, but as we have seen, internal insulating bodies are avoided for other reasons. Further, friction between surfaces in vacuum is greatly increased, probably by the removal of gas films.

For these reasons less and less use has been made of such guides in the progressive evolution of cathode systems, until they have finally been eliminated altogether, leaving a free-hanging system supported only on its seals. That a good standard of reliability is achieved by the use of such systems has now been proved in service by several years' operation of valves of the largest size, and by the similar good behaviour of their immediate predecessors.

Their immediately obvious disadvantages of relative lack of rigidity and of apparent relative fragility were of course borne in mind from the first, but the event has shown that these were by no means so serious as might have been feared.

The resultant magnetic force between the current in one limb of the system and those in its fellow limbs depends on the method of connection of the limbs. This may be exemplified by a brief description of various types of cathode system which have been used. Currents and voltages will be quoted in round figures. In the earlier types up to C.A.T.5 there were two 10-volt loops carrying 50 or 75 amperes in series. This necessitated a silica spacing insulator, usually sliding on a central support, but in C.A.T.5 hanging freely from the two stout 1·3-mm. filament loops. In these systems the resultant magnetic forces are radial, and so small that the weight of the filament and insulator and the viscosity of the tungsten suffice to preserve the shape.

For smaller currents in the limbs, e.g. C.A.M.2 with 12½ amperes, C.A.M.3, C.A.M.4, and C.A.T.6 with 35 amperes, and for higher wattage C.A.T.9 with 50 amperes, the series connection would give voltages about twice the preferred value of 20 volts, so the two loops had to be connected in parallel. In C.A.M.2 the currents were so small and the limbs so short that the magnetic forces were still negligible, but in the others special steps proved to be necessary.

In such systems the resultant force is larger, and is more effective since the loops are twice as long. It acts slightly outwards of the circumferential direction, and any distortion tends to increase the force, not decrease it as in the series case. To counteract it, each limb is given an appreciable preliminary distortion in two ways. The ends of each loop are set 6 mm. apart, which is closer than the 10 mm. separation of the central parts, and again both loops are bent in their own planes so that the ends are 14 mm. apart. The resulting shape of each loop is that of the gunwale of a boat, with a rise to bow and stern. The tension required to keep the loops so shaped in equilibrium is applied partly by the weight of the sliding central support, and partly by a spring acting on its outer end and pushing it downwards. The resulting system is stable throughout life, though, if the spring is defective, distortion becomes apparent after a few hundred hours.

It will be noted that in the parallel connection the tips of the loops are at the same potential and require no spacing insulator. Now the advantages of the series connection can be recovered if good electrical connection is made between the tips of the loops, and the supply leads are forked and crossed so that similar polarities appear at opposite ends of a diameter. The currents then run opposite ways in alternate limbs, as in the series case, and also—which is a great advantage—points at the same level on alternate limbs are at the same potential, and can be connected by finer wires, the tensions developed in which can counteract the resultant magnetic forces.

This principle of construction can be extended to include any even number of limbs, and it is adopted in all the larger types of valve. Thus C.A.T.10 has 6 limbs with triangular bracing wires, C.A.T.12 the same number on a wider base, C.A.M.6 8 limbs with square bracers, C.A.M.5 12 with interlaced triangular bracers, C.A.T.14 16, and C.A.T.19 24, the two last-mentioned with interlaced square bracers. The thick molybdenum leads are blackened to assist dissipation of heat by radiation, and are sawn into the required number of segments, and these are then bent to a jig so that they interlace accurately when assembled. The filament ends are arc-welded directly into holes or slots in the ends of the leads. This gives a slightly brittle joint, but, if required as in C.A.M.2, this brittleness can be dealt with by providing a clamp between the arcing point and the body of the filament. After arcing at the support end, the limbs of the multiple systems are adjusted to come together symmetrically at the far end without strain, and are then arced either all together in one operation, or in succession to a flexible ring of sheet molybdenum. The cross-bracing wires are then fitted in small V bends in the limbs placed to receive them. The spacing of the bracers is determined by the weight of the filament system below each bracing level.

The resulting system should be free from strain, but any residual strains are removed by raising to incandescence in a neutral atmosphere.

Appreciable distortion in after life is very rare. The stability of the system against electrostatic forces was established by means of a model in which weighted chains were substituted for the filaments.*

The presence of the cross-bracers also prevents an effect which had been troublesome in the smaller systems. This is the curious fact that the natural frequency of transverse elastic vibration of a tungsten filament, whatever its diameter, having the length required for a 19-volt working potential-drop between rigidly fixed points, is always very close indeed to the 100-cycle frequency of the magnetic forces due to a 50-cycle a.c. heating current. This natural frequency varies over a range of 20 % between cold and working temperatures, and therefore a resonance can easily occur during heating up, and, if the voltage applied results in currents in excess of the normal passing through the cold and brittle filament, the consequent vibrations in the plane of the loop can be notably destructive, breaking the loop at the tip. In the design of C.A.T.9, the natural frequency has had to be lowered to a safe value by making the ends of the supports themselves flexible in the plane of the vibration.

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* Le Rossignol and Hall, see Reference (2).

As in the case of the cathode seals there is no obvious difficulty in constructing still larger cathode systems on the free-hanging principle. Indeed, the mechanical stability is unaffected as the diameter increases, so that once the narrowest practical system, seen in C.A.T.10, was found satisfactory, the way was seen to be clear for larger systems.

Cathodes are still generally heated by direct current, but if more than two seals are provided, which on the present systems can readily be done, polyphase alternating current can be used, and in particular with four seals the hum-reducing connection using two phases in quadrature or otherwise.

(k) Cooling of anode

The rate of transfer of heat from the anode surface to the cooling fluid is so high, of the order of 1 kW per sq. in., that the actual mechanism of the process may be peculiar to valves. This has been investigated.* Space does not permit description of the work, but the results may be briefly summarized. The surface temperature of the anode is found by an embedded thermocouple to reach the boiling point of the liquid at comparatively low power dissipations. Various oils were studied, as well as water. Thereafter it can rise very considerably, as much as 40 degrees C. in some cases, even at our usual somewhat conservative rated dissipations. Heat transfer at these high rates is evidently facilitated through the formation of small bubbles of vapour, or of dissolved air if present, which break up the stagnant film of liquid in contact with the anode. Further, there must be some transfer in the form of latent heat of condensation of the vapour.

These processes evidently provide a very effective convection of heat, quite apart from the general motion of the liquid over the surface. It is thus doubtful if spontaneous turbulent motion is really necessary, and it may be remarked that, according to Osborne Reynolds's laws, mere constriction of the space between anode and jacket does not increase the probability of turbulence for a given volume of fluid passing. It should reduce the thickness of the effectively stagnant film, and seems to do so at the lower dissipations, but at the higher it also definitely retards the break-up of the film by small bubbles. At the higher dissipations certainly the results show that constriction can be a disadvantage. This is further illustrated in the case of low μ modulator valves, in which the bombardment is focused on the anode in streaks, and, in consequence, the cooling conditions become more complicated. Here constriction of the jacket lowers the temperature only a little, and is accompanied by irregular flow, setting up severe vibration. Probably the flow is intermittently diverted from the hot streaks by formation of large bubbles trapped by the constriction. The focusing effect has other curious results, which show that it is the overriding factor in such valves. Thus, if the anode voltage be fixed and the kilowatts dissipated be reduced by increasing the grid bias, the streak temperature is practically unaffected over a range of some 6 : 1 in dissipation, sharper focusing compensating for reduction in dissipation. Again, increase in anode diameter can increase the streak temperature owing to sharper focusing; it does so in C.A.M.4 as compared with C.A.M.3. Between the streaks the temperature does not rise above the boiling point. These valves are very sensitive to the cleanliness of the anode surface. If hard water is used a temperature of 250° C. can be attained.

More recent work appears to show that a forced air blast can be substituted for liquid cooling in any type of valve operating, as is usually the case, somewhat below the rated dissipation for water-cooling. For this, the surface for heat transfer is very greatly increased by soldering to the anode either blocks of copper perforated by a great number of holes, or a very closely spaced set of radial fins. By this means the anode temperature can be kept down to 140° C., which, as we have seen, is a quite usual value for water-cooling.

EVACUATION—THE EQUILIBRIUM STATE

(a) General

The general principle followed in all exhaust schedules is to subject each part of the valve for a sufficient time while it is on the pump to conditions more severe than it will meet with in operation. Thus the filament is overburnt, at the cost it may be of a few hundred hours of life, the grid is heated by bombardment to just short of the point where appreciable evaporation from the hotter parts would occur, and the anode is overheated with or without water-cooling—and we have already seen that even with water-cooling there can be considerable overheating given sufficient overload. The anode is not necessarily bombarded at all.

In the types which have bright emitting filaments no getter is used; as we shall shortly show, it is not needed. Independent evidence that this procedure has a sound basis is provided by the observed fact that gas evolution from heated metals and also from glasses is enormously reduced by a relatively small lowering of the temperature.

In addition to higher dissipations, higher voltages are applied in order to accelerate the fatigue of any tendency to flash-arc breakdown in the insulating properties of the vacuum, there being, as we have seen, no other insulation which is under any serious stress.

The schedules are judged by their results, and, where numbers permit, statistically, and in the light of reports of performance in operation.

This kind of procedure is necessary if the reliability of the product is not to be jeopardized by changing the schedules in ways which may prove to be ill-advised. Clearly, however, the further it is removed from the plane of trial and error by increasing insight into what is actually happening in the valve in manufacture and in operation, the more quickly will improvements be made and faults eliminated.

The true nature of the accepted gas test and the general behaviour of residual gas in valves with substantially cold anodes thus came in for investigation, and in both cases the results were somewhat surprising. Naturally, what was happening to the gas was much better understood when the test had been studied, so we shall take the work on the test first.

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* Le Rossignol and Hall, see Reference (2), p. 9, Fig. 12.

(b) The gas test and its meaning

The discriminating test is the now well-known backlash or gas test, in which the reverse grid current or "backlash" is measured under standard conditions of anode voltage and dissipation, the grid being necessarily negative. The backlash ratio is defined as the ratio of reverse grid current to anode current, and is usually expressed as microamperes per ampere. At a relatively late stage in the development of standard methods of exhaust it became clear that for hard valves the interpretation of the backlash test was more complex than was at first supposed. The simple ionization gauge law does not apply, i.e. the backlash ratio is not directly proportional to the equilibrium pressure. The factors which led to this conclusion are stated and discussed later, but to simplify the discussion a brief summary of the explanation is given at this stage.

In order to account for the observed behaviour of the backlash ratio r, it is necessary to assume that the reverse grid current contains two components: (1) the positive ion component r_p obeying the ionization gauge law which is of major importance for soft valves, and (2) a component r_0 of unknown origin which is of major importance for hard valves, where $r = r_p + r_0$. The experimental investigation was concerned, therefore, with the nature of r_0. Analysis of various test results had shown that r_0 continues to exist at substantially zero pressure; the residual grid current cannot, therefore, be a positive ion current, and must be an emission of electrons from the grid.

Consideration of the mechanism of this emission proved it to be photo-electric in origin, and caused by exposure of the grid to soft X-rays produced at the anode surface.

The principal features of the test results which the investigation had to explain, were as follows:—

First statistically, i.e. from large numbers of valves:—

(1) In a series of valves of identical construction the backlash ratio is never observed to fall below a certain limiting value. This needs little explanation on the photo-electric theory, because if gas were entirely absent, the backlash ratio would reach the limiting value r_0.

(2) Of freshly pumped valves, the majority show backlash ratios near to the limit, and few stray upwards to as much as 50 % more. This spread is unlikely to be due entirely to variations in gas pressure, owing to the enormous clean-up capability of water-cooled valves (described later). It is due only partly to slight changes in gas pressure, but partly to variations in the photo-electric component caused by adsorbed gas or traces of other impurities on the electrode surfaces.

(3) For valves differing in little besides the material of the grid, the limit is distinctly lower for molybdenum grids than for tungsten grids. For any type of valve r_0 for molybdenum is about 70 % of r_0 for tungsten. No satisfactory explanation could be found for this observation before the photo-electric theory was advanced, unless it was assumed that the residual gas pressure was actually lower in molybdenum grid valves, owing to gettering by the more volatile molybdenum. Although the residual pressure in a well-pumped valve is close to the limit of measurement, such a difference, if it exists, should have been detected by ionization gauge measurements. However, the photo-electric theory provided the explanation, as it was found that under the action of soft X-rays from copper the photo-electric emission from molybdenum is about 70 % of that from tungsten.

(4) For valves differing only in the diameter of the anode, the limit is lower the greater the anode diameter.

It was this last observation, more than any other, which led to a detailed investigation of the real meaning of the backlash test. For, if the reverse current is due entirely to gas, one would expect the backlash ratio to be greater in large anode valves because the electron paths are longer, and therefore more positive ions are produced. It was thought at first that for some reason the equilibrium pressure might be lower in large anode valves; and to decide the point a double valve having identical, but separate, grid-filament systems, and anodes of different diameter, was assembled inside a common envelope. This valve was known as the "Siamese Twin," and took the form of two valves with their anodes joined end to end, so that the communicating hole had the diameter of the smaller anode. There could thus be no question of a difference in pressure. Now when this double valve was hard, it was found that the large anode valve showed the lower backlash ratio. If the gas pressure was increased, however, the behaviour was as expected, and the backlash ratio increased more rapidly in the large anode valve, which eventually gave the higher backlash ratio. The pressure at which the backlash ratios were equal was about 2×10^{-5} mm. of mercury for anode diameters of $1\frac{3}{4}$ in. and 3 in. This result demonstrates clearly that for well-pumped valves gas ionization contributes little towards the reverse grid current.

Now, on the photo-electric theory the explanation is quite simple. As the anode is moved further away from the grid, the X-ray intensity at the grid surface and, consequently, the photo-electric emission fall off—inversely as the anode diameter for a circular cylindrical system. This is in good agreement with observations on different types of valve where anode diameter is the major variable.

Secondly, in addition to the above information from finished valves, we have, as a result of experiment:—

(*a*) For valves near the limit, the reverse grid current is proportional to the anode current at constant anode voltage, i.e. the backlash ratio shows little or no variation with the anode dissipation. This, of course, is true whether the grid current consists entirely of positive ions or of photo-electrons. It is only in very soft valves that the backlash ratio is appreciably affected by anode dissipation.

(*b*) For valves which are not fully bombarded, the backlash ratio is higher than for fully-pumped valves, although the residual gas pressure may be the same for both. A separate X-ray photo-electric experiment showed that for constant conditions of X-ray generation the number of photo-electrons liberated depended on the state of the molybdenum and tungsten surfaces. For surfaces which are treated in the same manner as grids for cooled-anode valves, adsorbed gas produces an increase in the photo-electric emission.

(*c*) For valves near the limit, the backlash ratio varies in a characteristic manner with the anode voltage (see Fig. 2).* This observation was at first as mysterious as the variation of backlash ratio with anode diameter. According to all experiments on ionization by collision, an increase of anode voltage in the range of voltages used should have resulted in the production of rather fewer positive ions, and, consequently, in a lower backlash ratio. However, the photo-electric theory disposed of the mystery, for reference to Fig. 4 will show that the curve relating photo-electric emission from molybdenum with X-ray exciting voltage is very similar to the backlash ratio-anode voltage curve for a molybdenum grid valve. There can be little doubt, therefore, that the increase of backlash ratio as the anode voltage is raised is caused by an increase in the photo-electric component of reverse grid current due to the increased intensity of the X-radiation.

It will be seen that the photo-electric explanation adequately fits all the previously anomalous observations. There has, of course, been no direct confirmation in an actual valve; such a test would necessitate screening the grid from X-rays, which is quite impracticable. However, the evidence in favour of this explanation is so complete that no doubt exists of its accuracy.

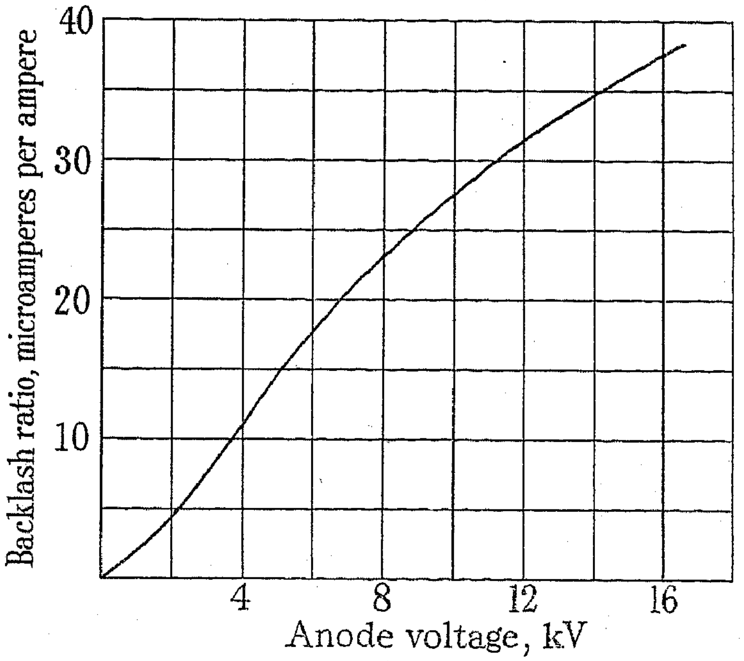

Fig. 2.—Variation of backlash ratio with anode voltage for a hard valve, type C.A.M.2 (molybdenum grid).

Other possible explanations were considered before the photo-electric theory was finally accepted. Various possible causes of grid current exist, including such factors as leakage across insulators, positive ion emission from the filament, and primary electron emission from the grid, but in water-cooled valves these factors, if present at all, are of negligible magnitude. Besides, the observed dependence of the residual reverse grid current on anode diameter and anode voltage narrowed the field of the investigation, and reduced the possible explanations to two:—

(1) The photo-electric explanation already discussed, and (2) an emission of electrons from the grid due to bombardment by high-velocity positive ions produced on the anode surface. This second hypothesis was also investigated experimentally in a valve having an extra grid placed close to the anode, by means of which the field at the anode surface could be reversed, thus preventing the escape of any positive ions formed there. The result of this experiment was entirely negative.

Photo-electric emission of electrons from tungsten and molybdenum under the action of X-rays from copper. It is impossible, in an actual valve, to separate the photo-electric component of grid current from the gas component. The apparatus shown pictorially in Fig. 3 (see Plate 2) was therefore built to study the emission of electrons from grid materials, i.e. tungsten and molybdenum, under the action of soft X-rays from copper.*

The radiograph is self-explanatory, and it will be sufficient to mention that the X-ray tube filament corresponds to the valve filament, the copper anti-cathode to the valve anode, and the photo-electric plates to the grid, either tungsten or molybdenum. Hence the X-ray tube voltage is analogous to anode voltage, the X-ray tube current to anode current, and the photo-electric emission to grid current. The ratio of photo-electric emission to X-ray tube current is therefore analogous to the residual backlash ratio.

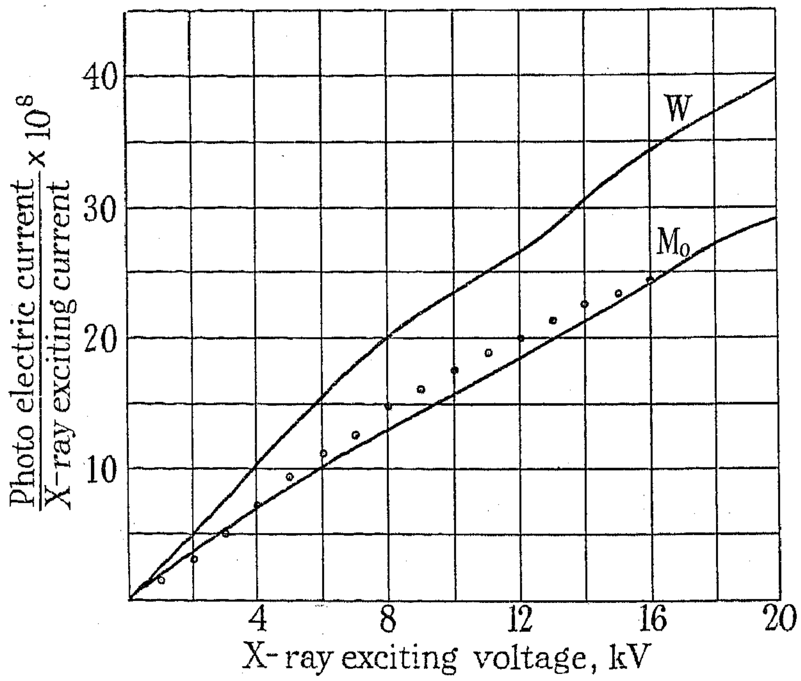

Fig. 4.—Photo-electric emission from clean tungsten and molybdenum under the action of soft X-rays from copper.

The points ⊙ are backlash ratio points, taken from Fig. 2 for a hard molybdenum grid valve (ordinates adjusted to fit at 16 kV).

Fig. 4 shows the result obtained after thorough degassing of the photo-electric plates. The increase of photo-electric emission as the X-ray exciting voltage is increased is due to the increased intensity of the soft radiation, and not to the appearance of progressively harder rays. This was proved by tests with an aluminium absorption screen. Tungsten shows a greater photo-electric emission than molybdenum, the ratio being approximately that previously observed in valves.

It will be seen from Fig. 4 that the qualitative agreement between the experimental results for molybdenum and a molybdenum grid valve is good, and the agreement

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* The pursuit of this investigation was encouraged by information from Mr. C. R. Burch, that curves similar to that of Fig. 2 were normally obtained from continuously-pumped valves of the Metropolitan-Vickers demountable type.

* See Reference (4).

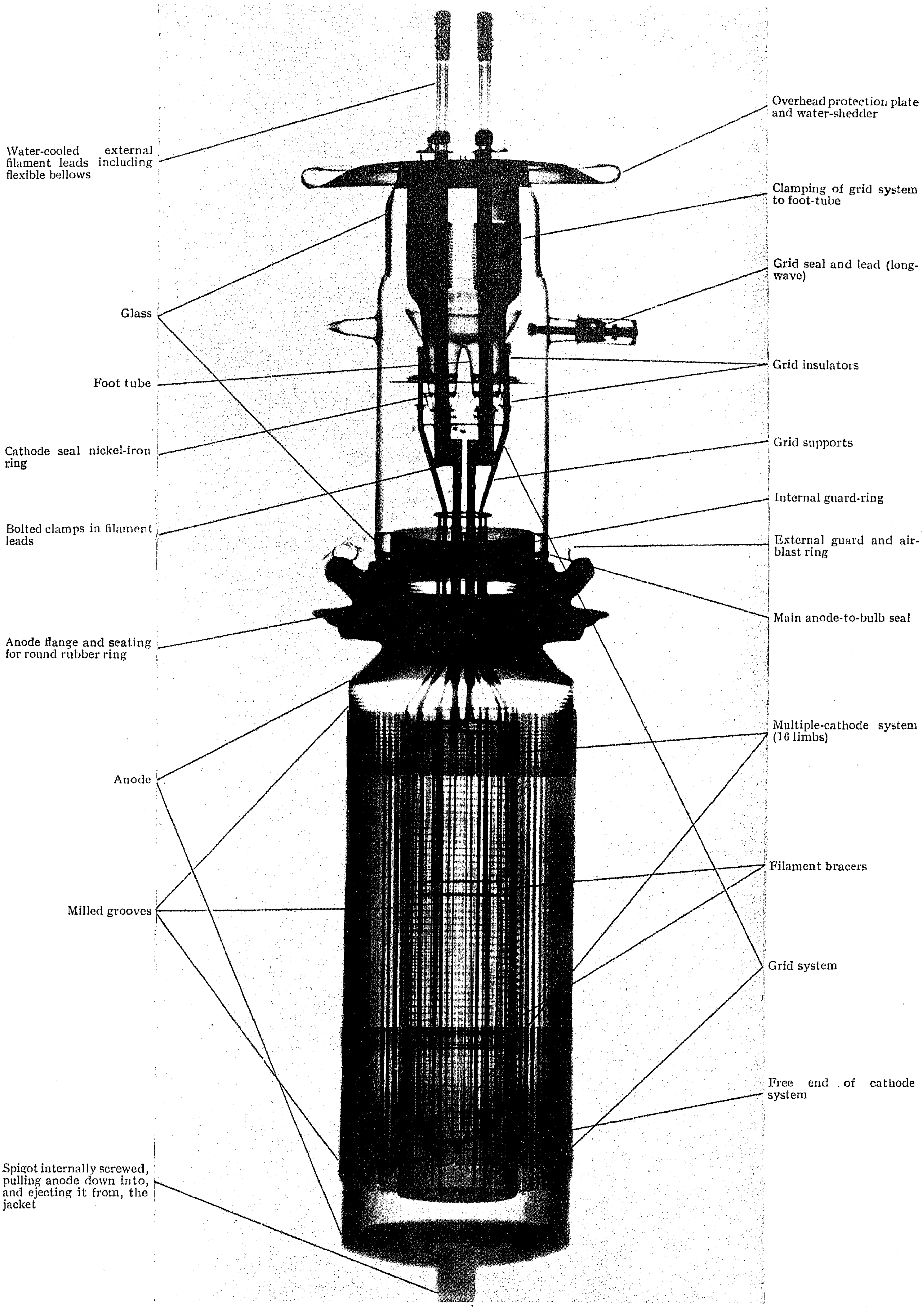
Water-cooled external filament leads including flexible bellows
Overhead protection plate and water-shedder
Clamping of grid system to foot-tube
Grid seal and lead (long-wave)
Glass
Grid insulators
Foot tube
Cathode seal nickel-iron ring
Grid supports
Bolted clamps in filament leads
Internal guard-ring
External guard and air-blast ring
Main anode-to-bulb seal
Anode flange and seating for round rubber ring
Multiple-cathode system (16 limbs)
Anode
Filament bracers
Milled grooves
Grid system
Free end of cathode system
Spigot internally screwed, pulling anode down into, and ejecting it from, the jacket

Fig. 1.—Radiograph of a typical high-power valve (type C.A.T.14). Approximately one-fifth actual size.

Plate 2

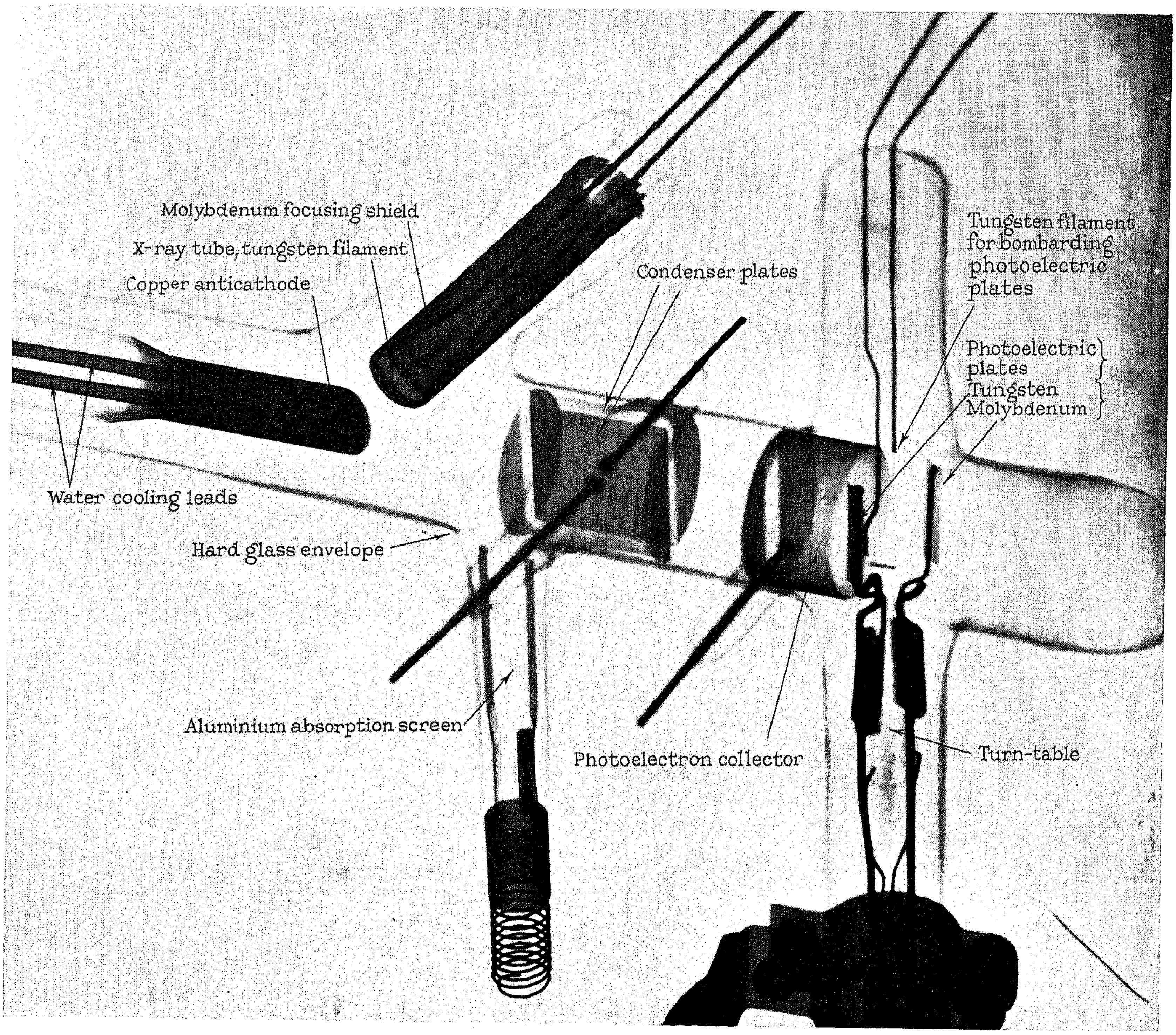

Fig. 3.—Radiograph of the apparatus used to study the photo-electric emission of electrons from tungsten and molybdenum under the action of X-rays from copper.

for tungsten is equally close. Quantitatively, the results obtained from the experiment just described may be used to calculate, by considering the relevant solid angles, the value of r_0 for a valve of any given geometry. Firstly, for a modulator valve having a parallel wire or " squirrel cage " molybdenum grid the experimental results predict a value of r_0 of 14×10^{-6} at 13 kV, whereas the observed value is 29×10^{-6}. Secondly, for a high μ tungsten-grid valve, the predicted value of r_0 is 28×10^{-6}, and the observed value is 45×10^{-6}. In both cases the calculated value is low, which may be due to an enhanced photo-electric emission in the valve caused by secondary X-rays.

The effect of adsorbed gas or other contamination on the molybdenum and tungsten plates was examined in the experimental apparatus. It was found that, after heat treatment corresponding to normal pumping of a valve, gas contamination increased the photo-electric emission, although other forms of contamination could reduce it. Hence, small changes of backlash ratio are not necessarily due to changes of equilibrium pressure, but may be due to traces of adsorbed gas or other contaminant on the electrode surfaces.

(c) Clean-up of free gas

A fully-exhausted water-cooled valve, with its large and very clean metal surfaces, forms a most effective pump and is capable of cleaning up surprisingly large quantities of free gas. Bearing in mind this large, but limited, clean-up capability, it becomes clear that one of the objects to be kept in view in developing exhaust schedules is the production of a valve capable of cleaning up, if required, the largest possible amount of gas liberated in operation. The clean-up capability may be measured in terms of the volume of gas cleaned up before the equilibrium pressure starts to rise.

The volume of gas required to exhaust the clean-up capability completely has not been determined, but for a valve having a volume of about 1 litre several cubic centimetres of air at normal temperature and pressure have been cleaned up under experimental conditions without by any means exhausting the clean-up capability. This corresponds to a reduction in pressure from 1 mm. of mercury to less than 10^{-6} mm. of mercury. Of course this degree of clean-up was not carried out in one stage. The experimental procedure, having first of all exhausted the valve in the normal manner, was to admit the gas in small doses, each dose being cleaned up before the next was admitted. Clean-up was produced by operating the valve under backlash test conditions which are known to be favourable.

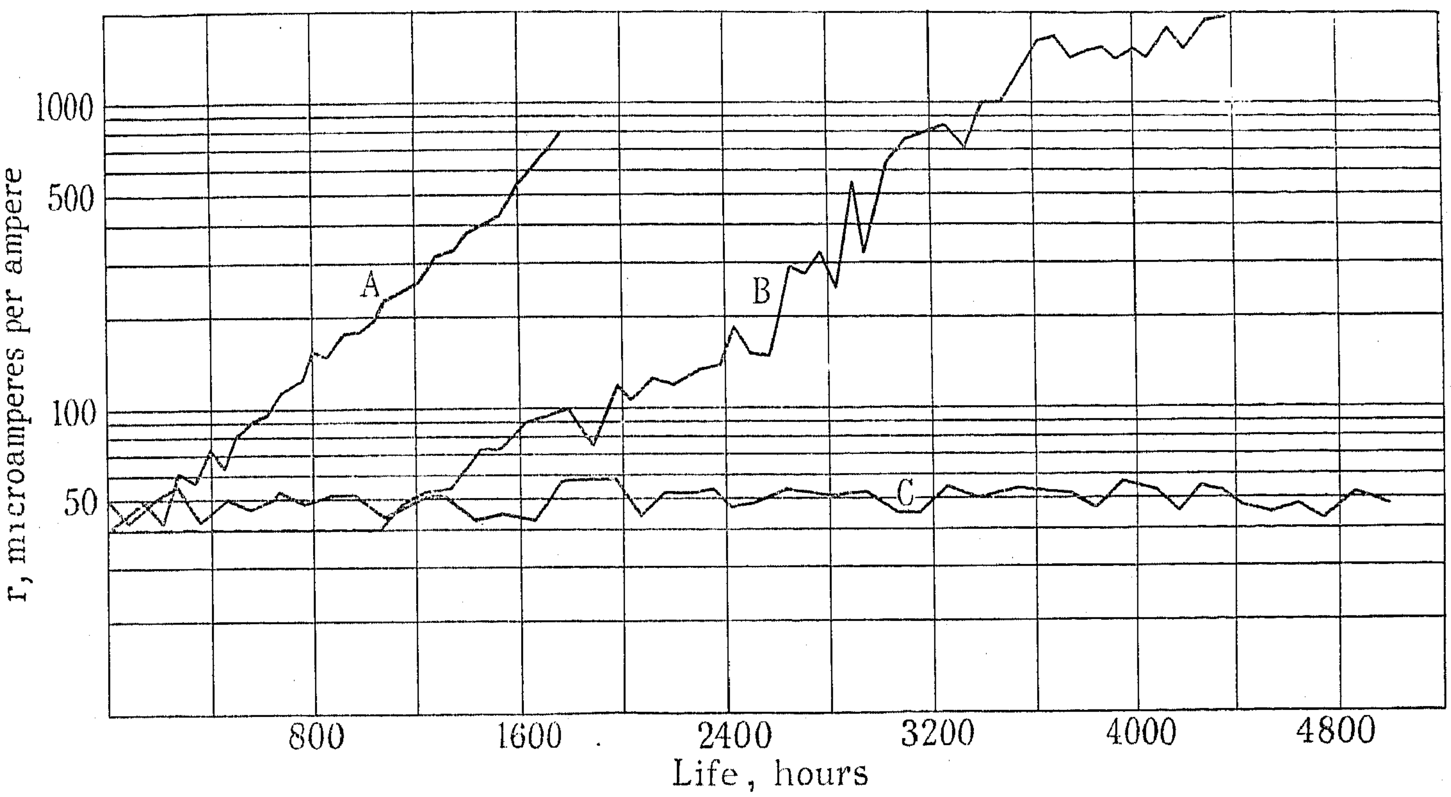

Fig. 5.—Variation of backlash ratio " r " for valves which soften during life.

Curve A. Valve with continuous source of gas and poor clean-up capability.
Curve B. Valve with continuous source of gas and greater clean-up capability.
Curve C. Valve with no serious source of gas and large clean-up capability.
Valve Type C.A.T.6.

This explains the recovery of valves accidently softened, for example through failure of the water supply. Cases have been recorded where such an accident caused the pressure to increase to the point where a glow discharge occurred, but operation of the valve under backlash test conditions with gradually increasing dissipation soon restored the vacuum to something approaching its original value.

Certain operating conditions are conducive to clean-up, notably those in which anode current flows when the grid is negative, at any rate for part of the operating cycle, e.g. normal " class B " or " class C " operation. Conversely, conditions in an absorber biased beyond cut-off when inactive, and passing anode current only when the anode potential is low and the grid is at a high positive potential, are not favourable to clean-up.

In practice this large clean-up capability renders the use of getters unnecessary. This is illustrated in Fig. 5, where the variation of backlash ratio during life is shown for valves which are now known to have had a continuous source of gas, namely a small piece of very slightly overheated mica. Curve A is for a valve with a poor clean-up capability, i.e. it is saturated when only a little gas has been absorbed, and the backlash ratio starts to rise early in life. The valve, however, would have been satisfactory had there been no source of gas. Curve B represents a valve having a greater clean-up capability, and the same source of gas as A. Here the backlash ratio is unaffected for 1 000 hours, but once the clean-up capability has been exhausted the pressure rises. Curve C is typical of present-day performance, where the clean-up capability is large and there is no serious continuous source of gas.

The smooth curve of Fig. 6 shows curve B of Fig. 5 replotted after subtracting the photoelectric component from the total backlash ratio, and it will be seen that the relation between time and the logarithm of pressure is linear over a considerable range. The straight portion is not always so long; but, nevertheless, it encouraged a search for a theoretical exponential relationship between pressure and time, but this could not be found on the assumption of a constant source of gas and clean-up by simple adsorption. However, this mechanism does account at least qualitatively for the general shape of the observed curve.

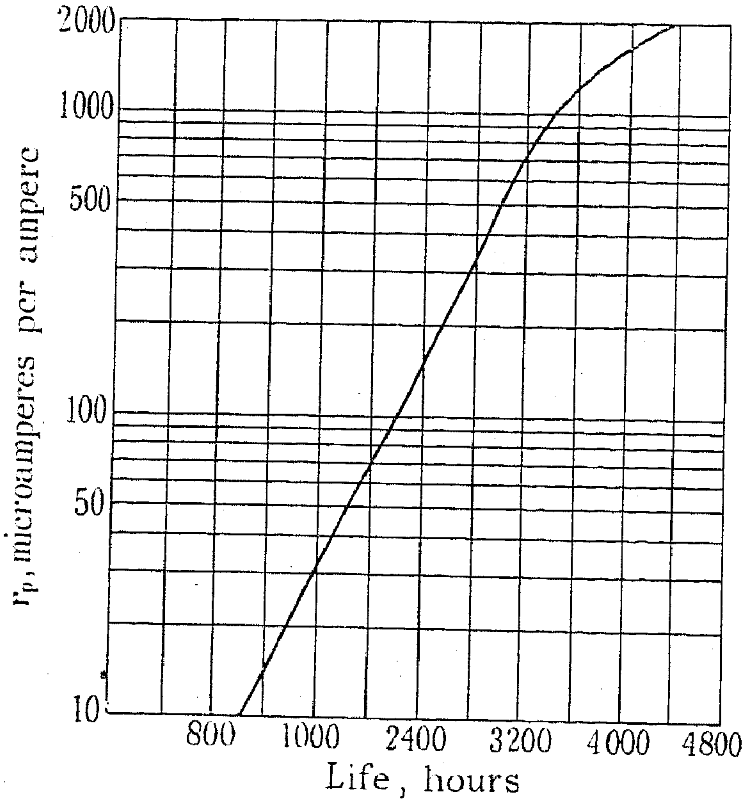

Fig. 6.—Variation of r_p, the gas component of backlash ratio, for a valve (type C.A.T.6) which softens during life. Curve B of Fig. 5 replotted after subtracting r_0, the photo-electric component.

CONTROL OF SATURATED EMISSION

(a) General

Although it is possible to obtain uniformity of performance between valves of a given type without much attention to the design and performance of their cathodes by providing a great reserve of emission so that no valve approaches thermionic saturation, such a procedure is expensive in respect of valve life, and particularly so in the case of bright-emitting tungsten filaments in which evaporation at emitting temperatures is appreciable. Means are therefore necessary for ensuring that valves of a given filament design, selected by the methods described by le Rossignol and Hall,* are operated as nearly as possible at the same emission. The regularity of performance attainable by taking full advantage of these means is illustrated in a later Section (Fig. 19.)

When emission data for individual valves are available, the rated values of filament voltage and current become quantities of secondary importance, subject to minor adjustments which are made when each valve is installed. The range of adjustment needed in practice has become less and less as the possession of emission data has reacted on the technique of filament manufacture.

It is impracticable in a routine test to take the full emission from the cathode continuously because of damage to the valve through overloading, except in the case of some low-impedance rectifiers, and also because the test may be misleading through overheating of the filament ends by the superposed space current, and so forth.

Hence, either the electrode voltages must be applied intermittently, or a reduced emission must be measured. The intermittent-voltage method has long been used for glass valves, the peaks of space-current being read on a peak-voltmeter connected across a resistor.

(b) The reduced-emission test

For high-power valves in which the full emission may be some tens of amperes the reduced emission method is preferred. The principle of this is to record readings of filament voltage and saturated emission, the values of emission increasing in roughly constant proportion over a total range of about 20 : 1, and then extrapolate to obtain the filament voltage required for the full rated emission. Since a graphical method of extrapolation is the most convenient, some way of plotting in which the observed points lie on a straight line which can be produced upwards has to be established.

For taking the readings the filament is heated with alternating current, since with direct current local overheating of the negative end by the superposed space-current may not be negligible. It can readily be shown that with a heating current H leaving the positive end of the filament and space-current of wave-form such that it has a mean value S_{mean} and an r.m.s. value $S_{r.m.s.}$ the equivalent heating current for producing the same local temperature in a non-emitting filament is

$$\surd(H^2 + S^2_{r.m.s.})$$

for alternating current, and

$$\surd(H^2 + 2S_{mean}H + S^2_{r.m.s.})$$

for direct current. The effect on the local emission can be estimated from the local overheating so defined.

If a more general treatment of the distribution of overheating is required, the same formulae will be found to apply to each elementary length of filament, provided that only the space-current emitted from the part on the positive side of the element is used; its value has to be obtained by integration, starting from the positive end.

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* See Reference (2), p. 2, Fig. 2.

Since even bright-emitting tungsten is susceptible to emission-poisoning at reduced temperatures, the filament is flashed at the full voltage before taking readings, and then the lower emissions are taken first. Further, the positive voltage is applied to all the other electrodes together. This is again to reduce the chance of poisoning, since in a water-cooled triode, for instance, a combination of high anode voltage and low grid voltage can produce poisoning, when a lower voltage applied to anode and grid does not. The control grid then takes up to about one-quarter of the total space-current, and if for the higher emissions the grid watts for full saturation are excessive, a lower positive voltage is used, and a small extrapolation to full saturation is made as described later in this Section.

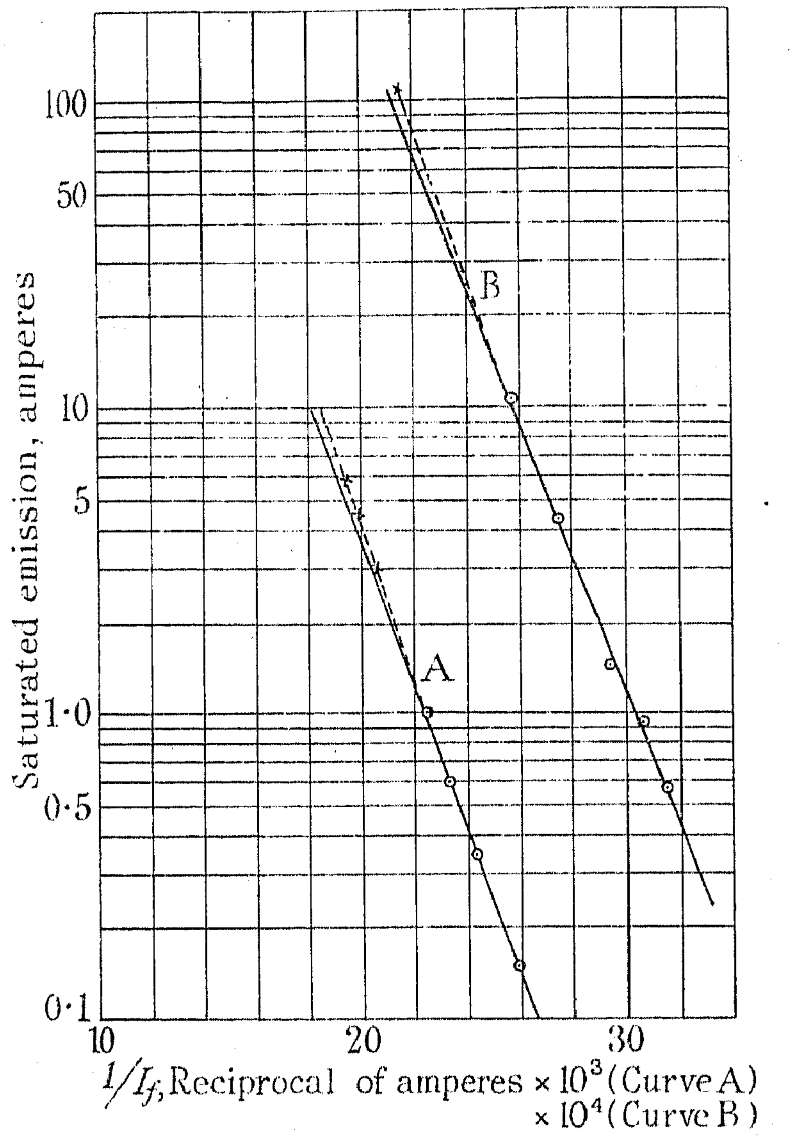

Fig. 7(a).—Extrapolation from reduced to full emission, simple method; log. of emission plotted against reciprocal of filament current.

Points ⊙. Normal test observations.
Points × (curve A). Additional observations for checking extrapolation.
Point × (curve B). Extrapolated point, using method of Fig. 7(*b*).
Curve A. Simple filament, valve type C.A.R.2.
Curve B. Multiple filament, valve type C.A.T.14.

(c) Extrapolation to full emission

Figs. 7(*a*) and 7(*b*) show two methods for the final extrapolation to full rated emission. Clearly the simple plot of log I_e against $1/I_f$ will only do for small extrapolations, although it is by far the best of such simple methods; whereas the method illustrated in Fig. 7(*b*) is much better, and this is, in fact, used for extrapolations up to as much as about 10 times the highest observed value.

The result of the extrapolation is to provide for each valve filament a " marked voltage " at which it gives a stated emission current which is 90 % of the value obtained from the plot, and is described as " 90 % saturated." This 90 % figure is adopted as being more useful to users, since the slow final approach to full saturation makes it unlikely that the 100 % value will ever be used in practice.

The plot is not necessarily made for every individual valve. In many types it is found sufficient to determine from a number of plots a standard extrapolation factor. For example in one well-known type, C.A.T.14, the "marked voltage " for 100 amperes 90 % saturated (111 amperes full saturation) is found to be 1·31 times the voltage observed for 10 amperes fully saturated, and the factor for that type is 1·31, which can be applied to a single test reading at 10 amperes emission. The variation of the factor for individual valves seldom exceeds ± 0·7 %.

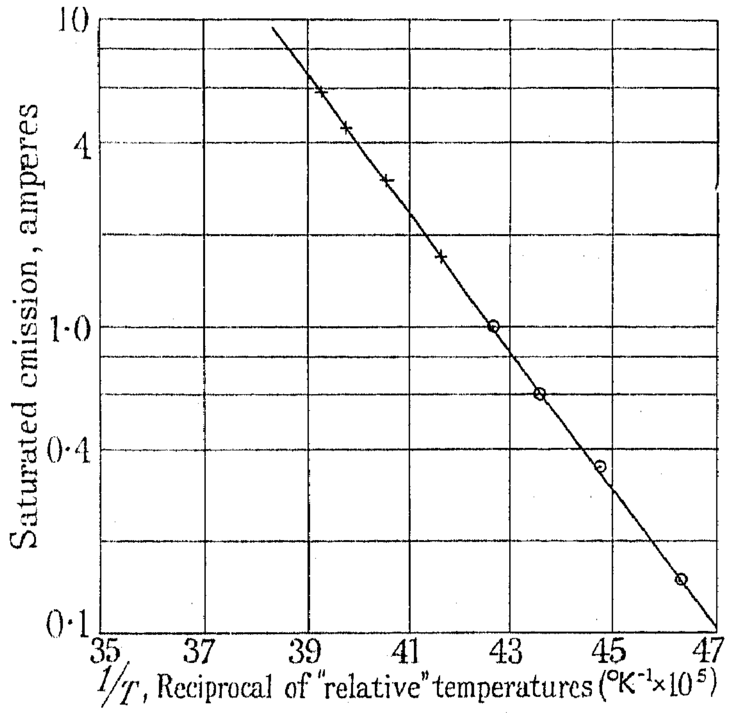

Fig. 7(b).—Extrapolation from reduced to full emission, standard method; log. of emission plotted against reciprocal of " relative " temperatures.

Points ⊙. Normal test observations.
Points ×. Additional observations for checking extrapolation.
Valve type C.A.R.2.

The method of plotting in Fig. 7(*b*) is based on the Richardson equation $I_e \propto T^2 e^{-b/T}$, but the T^2 term is negligible, and the T of the plot need not be a true temperature but can be anything whose reciprocal is a linear function of the true reciprocal temperature. The " relative " temperatures used for the plot are obtained by the following rapid approximation. The reading of filament current A is used and the corresponding relative temperature T_1 for a filament of length l_f and diameter d_f is found from Langmuir and Jones's* tables using the relation:

$$A = A'd_f^{\frac{3}{2}};\ A' = \sqrt{(W'/R')}$$

* See Reference (5).

where A', W', and R', are the heating current, power dissipation, and resistance of 1 cm. length in a filament of 1 cm. diameter at temperature T_1.

This relative temperature T_1 then corresponds to current A, voltage V_1, and emission I_{e1}. Using Langmuir's second relation:

$$V = V'\frac{l_f}{d_f^{\frac{1}{2}}}, \quad \text{where} \quad V' = \sqrt{(W'R')}$$

the other temperatures T_2, T_3, etc., corresponding to the other voltages V_2, V_3, etc., are found, and the plot is made.

The series of voltmeter readings V_1, V_2, etc., is preferred to a series of ammeter readings A_1, A_2, etc., because the combination of voltmeter and resistor is considered to be more permanently reliable than that of ammeter and current transformer, and because even a slight error in converting from amperes to volts after extrapolation might be serious.

Other similar methods have been examined, but the one here followed is found to be the most accurate. Since the voltage readings are actual terminal voltages, this process ignores corrections for graded cooling at the ends and supports of the filament, but these corrections are known to vary little over the range of temperature in question. This is one reason why the temperatures plotted are only "relative." Another and more fundamental reason is that the resistance property R' of the actual filament may not be the same as Langmuir's, nor the radiation property W' of the actual filament in its actual surroundings, which may vary in temperature and in reflecting power. Nevertheless, this will not matter so long as these properties preserve a due proportion over the range of temperature. There are two checks on the process available, one is to verify it directly with a plot of readings taken all the way to full emission without extrapolation, using a low impedance diode such as a standard water-cooled rectifier. This has given a very satisfactory confirmation, as illustrated by the points X in Fig. 7(*b*).

The other check is to calculate from the slope of the plot the value of the coefficient corresponding to the *b* of Richardson's equation, and it is found that when this is done the values of *b* obtained agree closely with the accepted value of *b* for pure tungsten.

(d) Emission statistics

The results of the reduced emission test for the first 61 valves to which it was applied are shown in Fig. 8, the position of each dot representing the amperes and volts required to produce the standard emission current, 1·0A 90 % saturated. The extreme range of variation is ±5 % in volts and amperes, i.e. ± 10 % in watts.

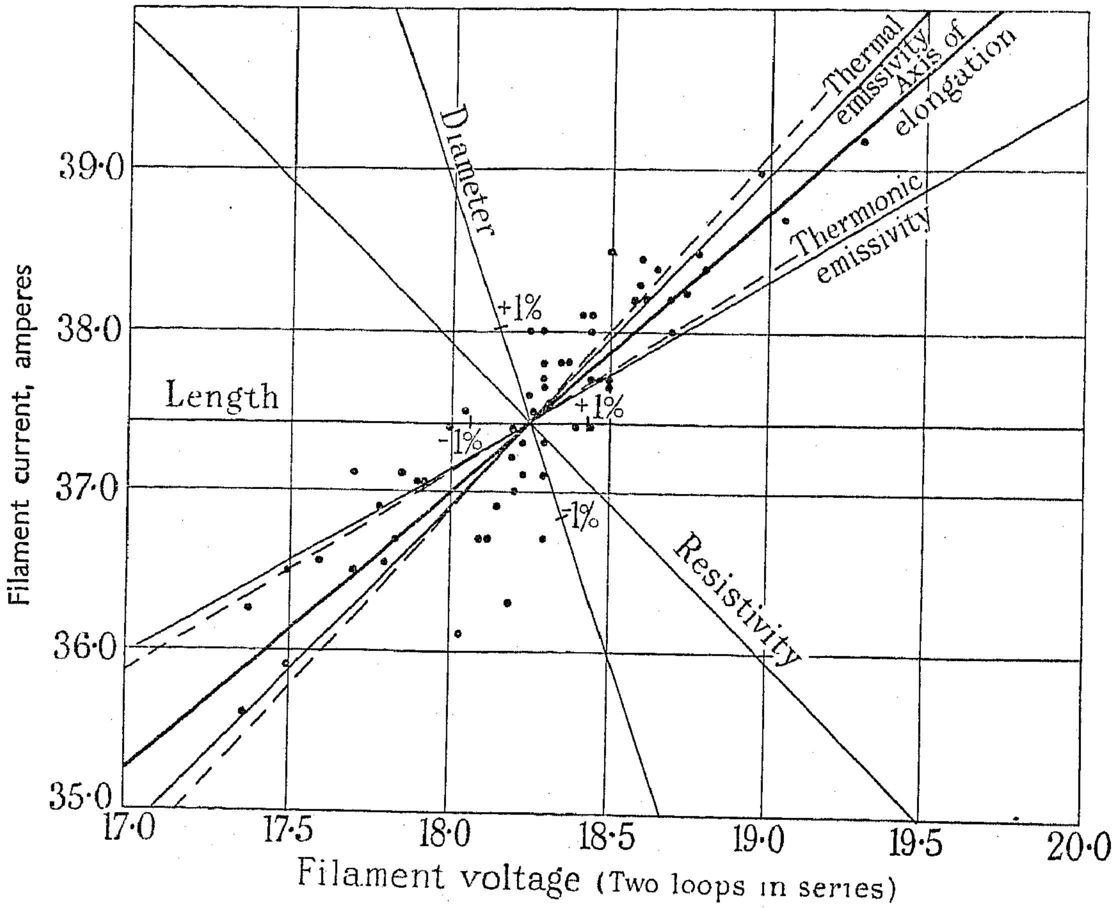

Fig. 8.—Reduced emission test, target diagram, showing axes of elongation expected from variation of each quantity (as marked) alone. Early valves (type V.T.26).

The marked elongation of the pattern indicates either that the observed variations had only one physical cause, or that if there were more than one cause then the several causes must be such as to produce closely similar effects. The axis of the elongation has been determined by the method of least squares; if a larger number of observations had been available, the direction of the axis might have been a little different, but the chances are 20 to 1 against it ever, for any batch of similar size, lying outside the broken lines, on either side of the axis shown.* Each of the other "axis lines" indicated on the diagram shows to a close approximation the direction which the elongation would have taken if only one physical quantity, that marked on the axis, had been subject to variation. Thus, to take the simplest case, variation in length will affect the voltage, but not the current, required to bring the filament to a given temperature. To reproduce the standard emission there would have to be a readjustment of the temperature to compensate for the change of area. Since, however, the emission varies very rapidly, this readjustment would be very small, so that the " length axis " will be a substantially horizontal line, as shown.

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* See Reference (6).

The direction of the actual elongation lies very close to the axis of thermal emissivity, that is, of the variation in watts required for filaments of the same dimensions and resistivity owing to variations in their thermal radiation to their surroundings. Under these conditions the resistance is constant, and the current and voltage will increase or decrease together and to the same extent, as the direction of this axis shows.

It is interesting to note what would be the effect on the diagram of variations in length and diameter, separately or together. These filaments were actually some 230 mm. long, so that variation in length exceeding 1 % must be regarded as very unlikely. The drawing of the wire is not, however, so easy to control, and variations of this order in diameter are much more probable.

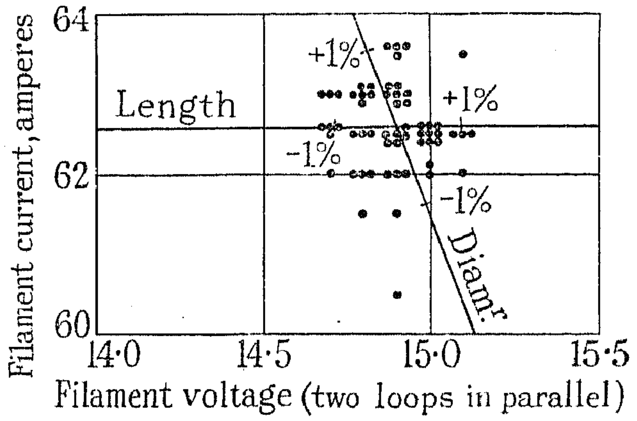

Fig. 9.—Reduced emission test, target diagram. Recent valves (type C.A.T.6).

Variations in length alone would thus produce only a very small elongation on the scale of the diagram; but variations in diameter would have a perceptible effect. Variations of both together would produce a small elongation in the direction of the diameter axis, but with some lateral spread. The point of interest is that a marked elongation in the direction actually observed could not result from independent variations of length and diameter, though the lateral spread could very well do so.

For such an elongation to result from variations in dimensions there would have to be a definite connection between the two, in the sense that all thin filaments would have to be short and all thick filaments long, and to a most unlikely extent. Since the two dimensions are under completely separate control this must be considered impossible.

The axes of thermal and thermionic emissivity are too close together for any definite distinction between these two causes to be made on the evidence shown in the diagram. All that could be said is that thermal emissivity is the more likely cause.

However, thermionic emissivity should be sensitive to heat treatment of any filament between repeated tests. No marked effects of that sort were observed, so that variations in thermal emissivity remained as the most likely cause of the observed elongation.

This conclusion indicated that the technique of wire drawing as affecting the final state of the surface of the filament required attention, and steps taken to ensure that filaments had a smooth bright surface resulted in a distinct and maintained improvement,* as exemplified in Fig. 9, showing the results of reduced emission tests on representative later valves of similar filament construction. The larger valves with multiple-filament systems containing more filament limbs must be expected to give still more uniform results, since these give an average of the performance of the individual limbs.

The improvement between the valves of Fig. 9 over those of Fig. 8 illustrates the value of a suitably designed and easily performed test on an isolated property of the product, once the meaning of the test has been sufficiently analysed.

SPACE-CURRENT CHARACTERISTICS IN THE OPERATING REGION

(a) General

The published characteristic curves for each type of valve are used for the purpose of choosing which type is best suited for a particular use. They are not apparently often used for predetermining refined details of performance such, for instance, as the percentage of modulation harmonics. Such predeterminations would require extreme precision in all parts of the curves, exact knowledge of circuit constants, and a good deal of labour in computation; in consequence, experimental study of optimum operating conditions is generally preferred.

Since valves which are expected to give a good life are operated with the filament temperature as low as possible, it is to be expected that there will be an approach to saturation in the part of the operating cycle where the space-current is largest. Now it is just where the space-current is largest that the valve is doing most of its work. Hence, it is particularly necessary that what happens in the region of the onset of saturation should be known, and together with this the effects of changes in the level of saturated emission.

To get characteristics directly by static readings is frequently just possible without undue risk to the valve, but high-power plant is needed. Further, static characteristics are liable to be misleading whenever the space-current is a sufficiently large fraction of the filament heating-current to influence appreciably the emission level—through ohmic overheating of the ends of the filament and " emission cooling " there and elsewhere.

Dynamic methods using peak-recorders or cathode-ray oscillographs are safer for the valve, but are liable to require even more plant. So a good many years ago, before cathode-ray oscillographs such as we have now were available, a simple method was devised by which space-current characteristics could be extrapolated from low-power readings, and this still seems to be adequate.

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* See Reference (7).

Simple extrapolation by extending the observed lower parts of curves using the 3/2 power law is not of much use. The onset of saturation is too gradual, more so than may be generally appreciated.

The index of the power-law curve, as shown by the slope of a doubly logarithmic plot, Fig. 10, begins to decrease at not much more than one-half of the saturated current, and to reach anything like saturation the voltage has to be raised to nearly twice its value at this point of departure. This early departure is probably valuable in practice, for it about doubles the length of the straight part of the characteristic; for present purposes, however, it is a complication. These effects are common to all types of valve, and seem largely independent of the geometry of their construction. The following analysis of the case of a geometrically simple valve can therefore be taken as being of general application.

This is the low-impedance rectifier type C.A.R.2, which has the " binocular " construction in which the filament consists of two straight limbs, each lying along the axis of a substantially complete cylindrical anode.* A doubly logarithmic plot of a characteristic of one of these valves is shown in Fig. 10.

Now in this kind of valve the symmetry of construction, the length (200 mm.) of the filament limbs, compared with their diameter (1 mm.), and the voltage-range of the characteristic, preclude explanation of this slow saturation to more than a very minor extent by most of the usual means, such as asymmetry of the collecting electrodes, screening of one part of the filament by another, and variation of potential, 20 volts (a.c.), along the filament itself. Some other reason or reasons have therefore to be found, and since the effect is but little greater in less-symmetrical valves as long as their filaments are straight, the reasons must be sought in the filament itself.

The most likely reason, physically, for the first departure is that there are local variations in the saturation value in different regions of the filament surface, and a calculation of the distribution of emission along the filament due to the end cooling, is found to account satisfactorily for the departure up to about 90 % of the saturation value, that is to say as far as anyone is likely to be practically interested in the characteristic. The calculation followed the methods of Forsythe and Worthing* and was verified qualitatively from the distribution of evaporation wastage in a long-life filament (see Fig. 17).

The method of extrapolation, to be described below, allows for any local variations in the saturated emission, whatever their cause.

It is, however, of interest to know the reason for the form of the uppermost part of the characteristic, since this might have some effect on the extrapolation. The

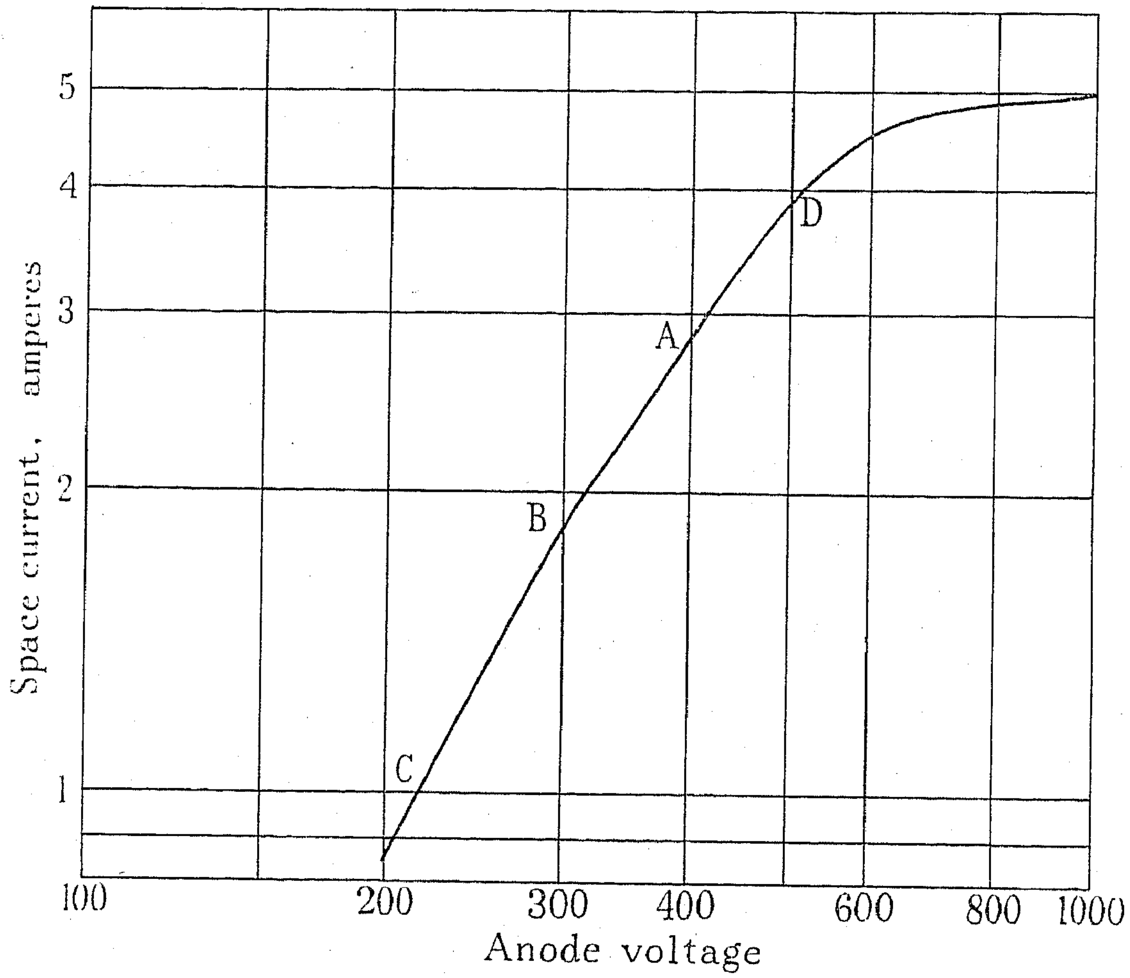

Fig. 10.—Characteristic (plotted to log. scales) of symmetrical diode (type C.A.R.2).

A. Departure from 3/2-power law.
BC. Incipient magnetron cut-off.
CD. Range of linear characteristic.

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* See Reference (2), p. 13, Fig. 16; and also Reference (17).

* See Reference (8).

explanation, according to some work by G. W. Warren, shortly to be published, is that we have here a kind of "telescoping" of the Schottky effect,[*] that is, the increase of the emission above the normal saturation value owing to a finite and increasing field at the cathode surface. That so large an increase as 10 % in the emission can be produced by the observed increase of collector voltage, arises briefly from the facts that as long as there is a fully developed space-charge there is no field at the cathode surface, and that as soon as the space-charge is reduced by the increasing initial acceleration of the emitted electrons, the full field due to the potential of the anode, or the equivalent equipotential in triodes, etc., is quickly established there. The initial growth of the field, and therefore of the emission, is found to be rapid, hence the "telescoping" mentioned above, and the increase of voltage required to produce what we have been calling saturation, which is really the establishment of the true Schottky rate of increase, agrees very well with what is observed in the C.A.R.2 type of valve, i.e. some 300 volts in Fig. 10. The extent of the effect is really very largely a property of the filament itself, and its existence accounts for the outstanding discrepancy in an earlier analysis of a diode characteristic by one of the present authors,[†] and takes priority over alternative explanations, e.g. in terms of local variations of the field at the cathode surface.[‡]

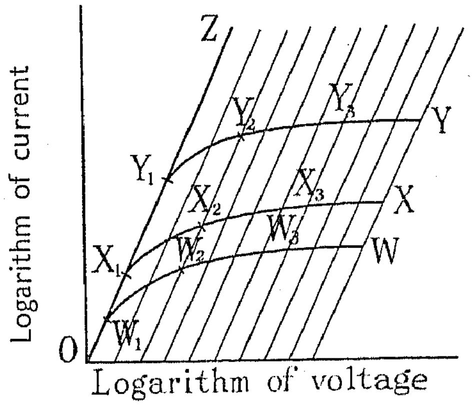

Fig. 11.—Diagram (plotted with logarithmic scales) illustrating extrapolation of space-current characteristic from lower to higher levels of emission.

After this analysis of the onset of saturation, we can now proceed to the method of extrapolation used. This employs a doubly logarithmic plot, logarithm of space-current against logarithm of electrode-voltage, and can be applied either to a diode, or to a triode with grid to cathode, or to a triode with grid bias, negative or positive, varied in constant proportion to the anode voltage, but with positive grid bias it is the space-current leaving the cathode that must be plotted.

If a number of curves for various levels of emission are plotted as in Fig. 11, the result at once suggests the method of extrapolation. We shall describe this in the first place as an empirical rule, and later discuss its probable theoretical basis.

The valve is usually connected as a diode, grid and anode together, so that only relatively low voltages and power are needed. One curve, OZ the "unsaturated" characteristic, is taken as high as grid dissipation permits, and at a filament voltage sufficient to ensure that there is no emission saturation, and another, OX, at the highest filament voltage consistent with reaching full saturation without exceeding the permissible grid dissipation. The other curve below OX is not essential, but is here included to illustrate the principle.

If lines parallel to OZ are drawn through points X_1, X_2, X_3, on OX cutting the lower curve OW in points W_1, W_2, W_3, the intercepts W_1X_1, W_2X_2, W_3X_3, are found to be very closely equal.

Hence to obtain a characteristic which saturates at a predetermined value Y, e.g. the saturation value at "marked volts," we simply extend the lines through X_1, X_2, X_3, parallel to OZ, and mark off Y_1, Y_2, Y_3, so that X_1Y_1, etc., are all equal to XY, i.e. to the difference in emission level. In practice OZ, except for very low currents, is usually very close to a straight line inclined at the 3/2-power angle.

If the valve had been connected as a triode, a curve would have been obtained which again is of the same form as OX, but displaced laterally to the right by a distance equal to $\log \dfrac{\mu + 1}{\mu[k + (1/\mu)]}$, where k is the factor of constant proportion between grid bias and anode voltage.

To calculate from OY the corresponding triode curves for various anode voltages, we have to replot OY, this time on ordinary paper, multiplying all voltage readings by $(\mu + 1)/\mu$, and then displace the curve so obtained by $-E_a/\mu$ successively in the ordinary way. The effects of changes of emission level, e.g. those due to changes of ± 1 volt on the filament, can be illustrated by a trifurcation of one of the curves.

The tail.

The lower parts of the curves towards cut-off are inserted from direct observation, since the dissipations are low. These lower parts do not follow the 3/2 power law, largely because of the superposed magnetron cut-off effect due to the field of the heating current, which may be between 30 and 75 amperes per limb. This effect is so strong that to get good readings near cut-off the filament current has to be held very constant. The displacement rule also may not hold here without modification, owing to progressive reduction of μ, as shown in Fig. 12, in which anode voltage is plotted against grid voltage for various anode currents. This form of diagram is convenient for extrapolating the straight lines of anode current to higher voltages than may be available. In types of valve which show a marked "tail" at the foot of a high-voltage characteristic, the lower lines on this diagram form a fan. However, in the C.A.M.3 type shown, and in other similar low-μ modulator valves, there is very little tail indeed. C.A.M.3 has four straight-filament limbs at the corners of a 10-mm. square within an octagon of grid wires 25 mm. in diagonal. The characteristics are the same whether the filaments lie behind four grid wires, or in the spaces between them. The absence of tail implies a high degree of uniformity of the field round the circumference of the filament in presence of the space charge, i.e. absence of *Inselbildung*.

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[*] See References (9) and (10).
[†] See Reference (11).
[‡] *Ibid.*, (12).

The addition of a fine spiral to the grid structure to increase the value of μ has little effect on the tail.

Division between anode and grid.

When the grid bias is positive, values of anode current have to be obtained by subtracting the corresponding grid currents from the total space-current. The grid currents are obtained by plotting the ratio of grid current to space-current against the ratio of grid voltage to anode voltage, as in Fig. 13. On such a plot we find that outside the region of secondary emission all the plotted points corresponding to various permissible currents and voltages lie on a straight line, and it is assumed that values of grid current otherwise unattainable can be obtained from this straight line. Theoretically we should use $(E_g/E_a)^{\frac{1}{2}}$ for the plot instead of E_g/E_a, since that would be the rule for the simple repulsion of particles from a charged wire. This makes little difference, but E_g/E_a is better, perhaps because of the known development of space-charge round the grid wires. The secondary emission region will be considered later in this section.

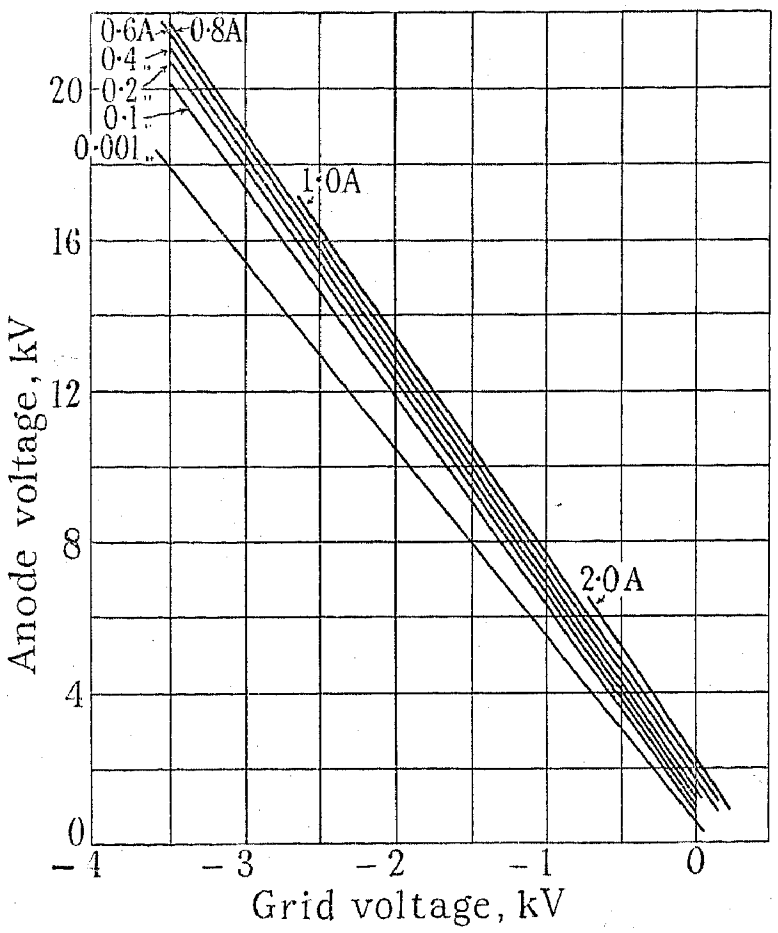

Fig. 12.—Current-contour characteristic of low-μ valve (type C.A.M.3), showing reduction of μ at minute currents only.

Theoretical.

Returning again to the main extrapolation on the doubly logarithmic plot, Fig. 11, the theoretical justification of the procedure can be referred to the equations of motion of the electrons, but can be explained as follows:—

At any point of the curved portion of a plot, parts of the filament are emission- or temperature-controlled owing apparently to the end-cooling, and parts voltage, i.e. space-charge, controlled. Suppose now that we have a curve, e.g. OX, plotted up to the limit of dissipation of the valve, and we want to draw a curve, such as OY, up to a larger current, say up to n times the current. If we increase any voltage $n^{2/3}$ times, the currents from all the voltage-controlled areas will be increased n times, if the temperature is simultaneously raised by not too small an amount. Again, if we increase the temperature by means of the heating current so that the saturated emission of all parts is increased n times, the temperature-controlled parts will give n times the former current, provided the voltage is increased sufficiently to keep them saturated. If now we do both those things together, e.g. pass from X_2 to Y_2 on the plot, those parts which were temperature-controlled will remain so, and those which were voltage-controlled will also remain so. For, consider a boundary line dividing the two states of the emitting surface. On the one side the necessary excess of voltage, and on the other side the necessary excess of temperature, are still maintained, so that the boundary line does not move. The two areas on either side are thus the same as before, and as the two simultaneous adjustments ensure that the current from each is increased n times, the total current is increased n times.

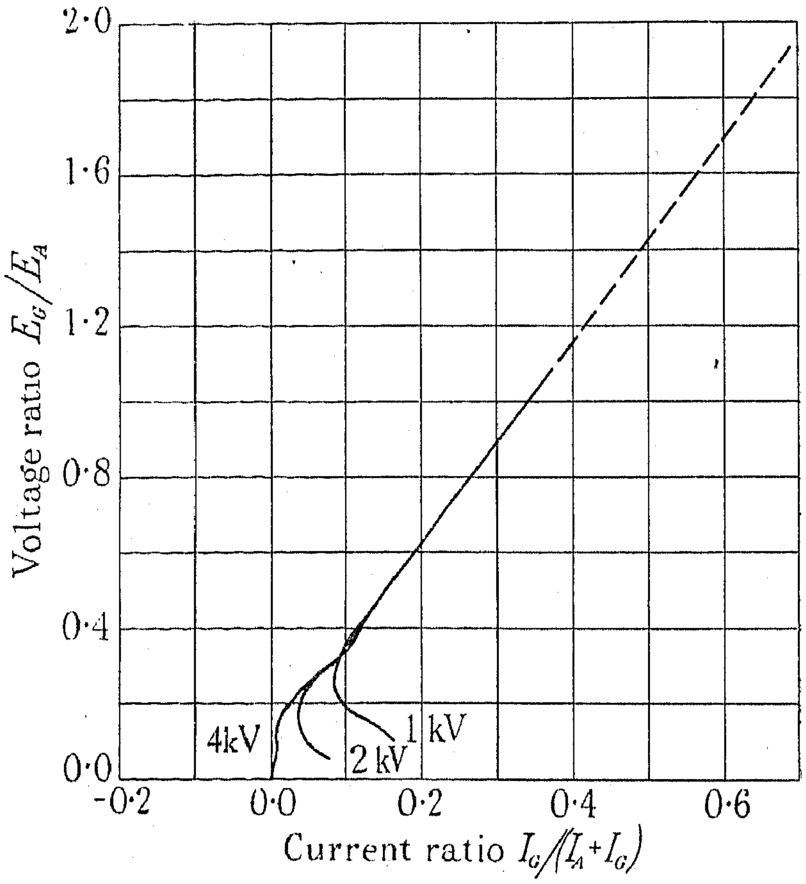

Fig. 13.—Division of space-current between anode and grid.

To sum up, we have arrived at the rule that from any observed characteristic any other can be obtained by increasing the temperature so as to produce an n times increase of saturated emission, and by increasing all controlling voltages $n^{2/3}$ times.

According to the argument, the rule can be applied without our having to know anything of the properties or circumstances of the various local areas of the filament, provided only that we feel justified, e.g. by comparing two curves such as OW and OX, in supposing that they do not appreciably interfere with one another. It does not apply in the region of the telescoped Schottky effect.

(b) Grid-current characteristics

Secondary emission.

The general form of the lower part of the grid-current characteristics is decided to a large extent by the amount of secondary emission present. The secondary-emission coefficient (number of secondary electrons produced per primary bombarding electron) depends on the bombarding voltage and on the state of the bombarded surface. In general, an electropositive contaminant increases the secondary emission, and an electronegative one reduces it. There is comparatively little variation from valve to valve of any type, which indicates that the grid of a high-power valve represents a very clean metal surface, and the secondary emission coefficient is approximately that of the pure metal. In the absence of any attempt to reduce the secondary emission, the type of grid-current characteristic often obtained is shown in Fig. 14 (lower broken curve). This can be dangerous for valves operated in circuits having high grid resistance, as the grid can reach and maintain a high positive potential, with consequent possibility of damage to valve and circuit components. Excessive secondary emission is therefore undesirable, and it became necessary to keep the magnitude of the effect under control. This can be accomplished in various ways. For example, the use of tantalum as a grid material results in low secondary emission, and the same result can be achieved by coating a tungsten or molybdenum grid with a substance, such as carbon, having a very low secondary emission coefficient. In fact it has been found that, if the internal surface of the anode is treated with carbon, sufficient finds its way on to the surface of the grid during pumping to produce a marked decrease in the secondary emission coefficient.

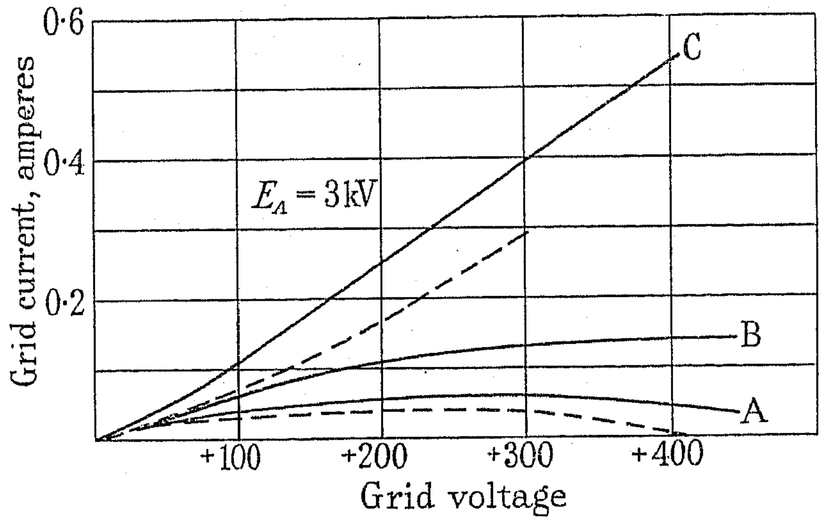

Fig. 14.—Control of the grid-current characteristic by movement of a grid band surrounding the cool ends of the filament.

The broken curves show the limits of valve-to-valve variation before control was applied.

Valve type C.A.T.6.

Apart, however, from the grid material, we have found that the secondary emission is not uniform along the length of the grid, but is much greater from the ends, which are opposite the cool ends of the filament, than from the central active region. Proof of this is obtained from the observation that dimming the filament results in enhanced secondary emission. This behaviour is fundamentally due to the temperature distribution along the length of the filament, and may be explained briefly as follows:—

Owing to the distribution of temperature along the filament, there is considerably more space charge around the grid wires in the central active region than at the ends. This causes the accelerating field tending to remove electrons from the grid to the anode to be stronger at the ends of the grid than in the active region. The ends of the grid therefore contribute a greater number of secondary electrons per unit length than does the active region. Now, if the ends of the grid instead of being an open structure are made in the form of a solid band surrounding the cool ends of the filament system, two things will happen. First, the ends of the grid collect more electrons, and, secondly, the secondary electrons, being produced on the inner surface of the band where the accelerating field of the anode is very weak, cannot escape, so that the effective secondary emission is reduced.

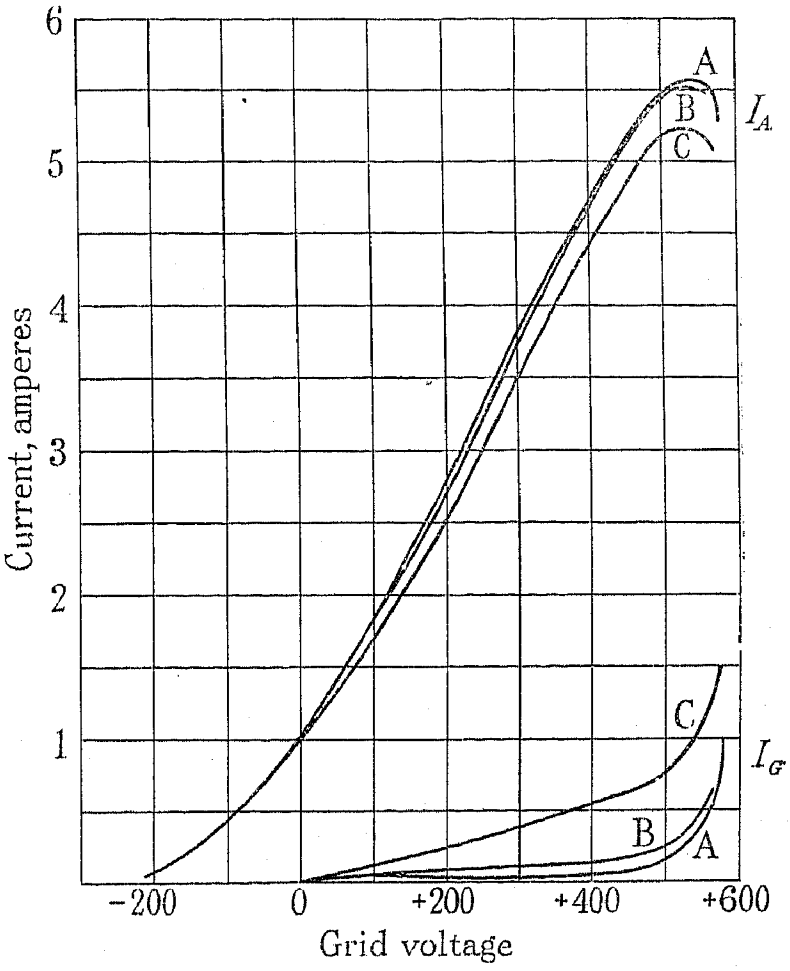

Fig. 15.—Effect on the "class B" operating characteristic of the distance by which the grid band overlaps the cool ends of the filament.

Curve A. Grid band overlap 2·5 mm.
Curve B. Grid band overlap 12 mm.
Curve C. Grid band overlap 20·5 mm.

Valve type C.A.T.6.

This effect was explored in a C.A.T.6 valve, having a movable grid band at one end of the grid, and the results will be seen in Fig. 14. Curve A shows the grid-current characteristic when the band overlaps the filament system by only 2·5 mm., the length of the filament loop being 180 mm. Curve B shows that when the overlap is 12 mm., the effective secondary emission is reduced considerably, and with an overlap of 20·5 mm. (Curve C) the secondary emission is entirely swamped, but here the band covers parts of the filament at nearly full temperature, and steals anode current. The effect of these changes on the class-B operating characteristic is shown in Fig. 15. For type C.A.T.6 a grid-band position corresponding to curve B has been adopted, and other types are treated on their merits.

One of the results of secondary emission is that the number of watts dissipated in the grid cannot, under most operating conditions, be easily calculated. Even under static conditions the product of grid voltage and grid current only gives the true number of watts if there is no secondary emission. In order to measure the grid watts any instrument which gives a measure of grid temperature may be used, and perhaps the simplest method is the use of a surface thermopile as a radiation indicator. Although the amount of radiation from the grid is usually much smaller than that from the filament the sensitivity of the instrument is sufficient to give an accuracy of, at worst, 5 %, which is ample in most practical cases.

Magnetic modification of grid characteristic.

The upper part of the grid characteristic depends mainly on the ratio in which the space-current is divided betweed grid and anode.

That there is another factor controlling this ratio, additional to the geometrical dispositions, is shown by the difference between C.A.T.14 and C.A.T.12.

In these two types the relative positions of filament wire, grid structure, and anode, differ very little, and yet the grid current is proportionately appreciably different in C.A.T.12.

The further factor seems to be the magnetic field of the filament heating-current, which is some 71 amperes per limb in C.A.T.12, and 56 amperes in C.A.T.14. The influence of this does not extend very far out, because in C.A.T.10 with the same 71 amperes filament placed only some 30 % further from the grid the current ratio falls to the same value as in C.A.T.14.

This effect has not been studied fundamentally in terms of the motion of the electrons; still there is some further experimental evidence about it. When oscillograms are taken with the filaments heated by 50-cycle alternating-current—though this has been done only with another type, C.A.T.6—the grid current is found to be strongly modulated at 100 and 200 cycles, having two maxima and one minimum in each half cycle of the heating current.

Hence even with d.c. heating we should be led to expect different ratios of grid current to anode current if the filament heating-current is different, as it is between C.A.T.12 and C.A.T.14.

This effect must also have some influence on the output modulation, or " hum," produced when amplifier filaments are heated by alternating current. If there is a grid-leak, as in Class " C " anode-modulated, there will be grid modulation. Again, whenever the grid current is comparable with the anode current the latter will be modulated by subtraction of the grid current from the total space-current. Finally, there may be direct modulation of the space-current itself. Where the resultant hum is small, it is possible that one type of modulation may be opposing the others.

In an experimental version of C.A.T.6 which had a three-phase filament system, even " static " oscillograms, i.e. these taken with fixed grid and anode voltage, showed very complicated wave-forms, consisting mainly of triple-multiple harmonics—300, 600, etc., cycles. The three-phase arrangement could not be expected to eliminate such harmonics, and it is found that a pair of standard single-phase valves with filaments heated in quadrature by Scott-connected transformers and eliminating both 100- and 300-cycle components gives a much lower level of hum.

FLASH-ARC BREAKDOWNS (ROCKY POINT EFFECT)

The spontaneous breakdown of the vacuum insulation leading to a complete short-circuit of the valve by a vacuum arc, and occurring after a time-lag of anything up to some thousands of hours, was described to this Institution some five years ago.* It was felt at that time that, although the initiation of the discharge was not understood, enough was known about its later stages to enable breakdowns in operation to be made so rare as to be negligible.

A residue of doubt whether this would be possible for larger valves was already being dispelled, although it was known that the same degree of protection as for smaller valves could not be obtained on the circuit side.

These tentative conclusions have been shown by the experience of later years to have been sound, in that a progressive increase in anode voltage up to 18 or 20 kV for the later types of valves has resulted in little or no trouble in six stations which have been in operation for one or two years, using two different types of valves.

The establishment of this later improvement has resulted mainly from a general increase in the smallest clearance between the grid and anode systems. This increase was adopted in valves intended for the higher voltages, after it had been found by direct comparison that an increase from 9 mm. to 25 mm. in the C.A.T.6 class of valve raised the permissible overvoltage on a short test by more than 50 % above its normal value for the smaller clearance.

Investigation on the lines of the previous study showed that the way in which increase of clearance reduces the chance of breakdown is not altogether simple.

By using a movable anode in the form of a copper disc mounted on a metal " bellows " opposite to a flat-headed loop of 1 mm. tungsten wire as cathode, two things were established.

First, as to the voltage required to produce a discharge with a short time-lag, there was the somewhat surprising result that with the electrodes in anything approaching their final state it was unusual for a reduction in clearance from 12 mm. to 3 mm. to produce any notable fall in the discharge voltage. This is illustrated in the upper pair of curves in Fig. 16. Here the surface field in kV per cm. on the side of the cathode loop facing the anode is shown for each discharge point.

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* See Reference (3).

The progressive increase of field towards the smaller spacings where a discharge should be easier to establish makes it seem unlikely that breakdown is any simple consequence of the field, high though that is. The results shown by Mason* for smaller spacings from 1·2 mm. downwards, agree better with our general experience than his stated conclusions would imply. His later points show a similar rise in field.

Anderson,† however, has published results generally similar to ours.

The points appear in pairs because the observations, starting from the largest spacing and returning one step back from the smallest, were taken alternately with a voltage pulsating at 700 kilocycles between zero and the value shown, and with a continuous voltage. The small difference in the run of the two sets of points shows that the time required to build up conditions for a breakdown is very short.

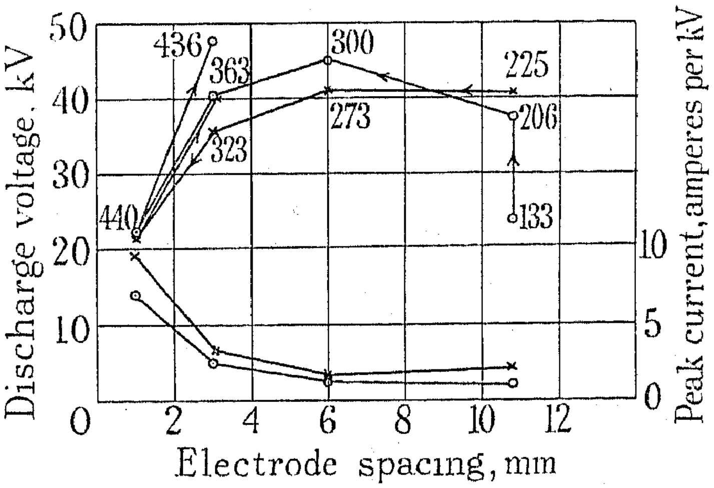

Fig. 16.—Flash-arc (Rocky Point) discharges in experimental tube with variable electrode-spacing.

Upper curves. Discharge voltage.
Lower curves. Peak current in discharge.
Points ⊙. Voltage pulsating at 700 kc.
Points ×. Voltage continuous.

Examination of the currents passing in the discharges —which have not so far been recorded by other investigators—throws some further light on what is happening.

The lower pair of curves show the peak value of the discharge current expressed in amperes per kilovolt. The theoretical value for an undamped discharge of the circuit used, 0·0002 μF and 4·5 μH is 7 amperes per kV and the observed peaks fall below this for the larger spacings, so much so that at 6 mm. or over at least nine-tenths of the initial energy of the condenser is used in building up the arc. Even so, in this particular experiment the current readings were higher than usual at the larger spacings.

Evidently the important point is that increase of spacing reduces the chance of a complete breakdown.

Nevertheless, there are other conditions to be satisfied, for if, as in Fig. 16, the electrodes were in a state where the general level of discharge voltage was below the maximum, change of spacing made much less difference to the currents, and so little or no advantage from the increased spacing remained.

However, the satisfactory operating experience shows that such conditions do not occur in finished valves.

The view that the presence of residual gas at relatively high pressures has no influence on the breakdown has been supported in operating experience. Considerable numbers of valves have been allowed to soften in operation up to pressures of the order of 5×10^{-4} mm. without any flash-arc breakdowns being reported.

The main principle on the circuit side of not permitting too much discharge energy to be too rapidly and continuously available still stands.

OPERATING CONDITIONS

About the operating conditions for high-power valves within their stated ratings there is little to be said. The need for constancy of filament supply voltage, if the proper average life is to be attained, is now well understood. For switching on the filaments and applying the anode potential, certain recommendations are made.

In the case of the filaments, the voltage should be applied in increasing steps over a period of about 30 seconds. The process is necessarily slow for two reasons. First, when the filament is cold its resistance is low, and the current it would take from a well-regulated supply at full voltage, e.g. a transformer tapping which happens to suit, might be 10 times the working value. The magnetic forces would thus be increased 100 times, and magnetic distortion would be unnecessarily encouraged, apart from possible catastrophic action on the initially brittle filament. Secondly, the temperature can overshoot the working value locally if there is sufficient local increase above the normal resistance per unit length, because of the local concentration of voltage in any region whose temperature, and therefore resistance, is rising more rapidly than elsewhere. The natural instability of the filament in this respect is thus exaggerated, and all filaments being affected more or less alike, the result may be simply that the average life is reduced without the reason being at all obvious. It is therefore recommended that for the first 10 seconds the filament current should not exceed 70 % of the working value; for this the applied voltage will have to be only some 7 % of the normal. After this, the voltage may be increased fairly rapidly, but so that the current never exceeds 130 % of the normal. The final voltage regulation has to be performed some minutes later when the leads, etc., will have reached a steady temperature.

As regards the anode potential, after any interruption during normal operation this may be switched on at a maximum of 70 % of the normal, and then after 2 seconds increased in one step to the full value.

For new valves, or valves which have been out of commission for some time, it is recommended that the initial voltage should be 5 000 for 10 minutes. This is followed by a gradual increase to the full value in not less than 20 minutes.

The cooling water must be pure enough to keep the anode free from serious deposits without requiring inconveniently frequent cleaning. For this it is recommended that the total dissolved solids shall not exceed 15 parts per 100 000, nor the solids precipitated on boiling 4 parts per 100 000, nor the conductivity 300 reciprocal megohms per cm. cube. These are limiting values; purer water is better.

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* See Reference (13).

† *Ibid.*, (14).

LIFE

The high standard of reliability demanded of transmitting valves to-day has already been referred to. Premature failures are to be avoided, and the failure zone restricted to a reasonable statistical percentage of the average life. A long average life is also desirable. Some idea of the standard achieved may be obtained from the overall average life of all reported failures of all types of cooled-anode valves. This has increased from 4 700 hours in 1932, to 6 400 hours in 1936.

There are many possible causes of failure, but filament burn-out is the only really fundamental cause. It may be delayed but cannot be prevented. The maximum life any valve can have is therefore the life of its filament, and we shall first consider the various factors which affect the life of a filament of pure tungsten, operating at a temperature of about 2 500° K. Filament failure is due to gradual wastage by evaporation, and eventual overheating of the thinnest point. It has been found, by examination of large numbers of burnt-out filaments, that failure occurs when the diameter is reduced by 10 % by evaporation. This rule holds whatever the initial diameter, and was first enunciated at the American General Electric Co.'s laboratories, for lamp filaments of 0·1 mm. diameter or less.* The wastage of various filaments of different original diameters between 0·82 and 1·33 mm. is shown in Fig. 17. There is a considerable length over which the wastage is uniform, indicating the region of uniform temperature. The ends show practically no wastage; this is due to cooling by the supports, and Fig. 17 also shows the cooling effect of the bracers. Evaporation appears to proceed normally until the diameter is reduced by about 10 % over a considerable proportion of the length; instability then sets in, and the temperature of a short length increases rapidly. Once started the effect is cumulative, and burn-out follows quickly. This point is usually, but not always, the expected hottest point of the filament. The 10 % reduction in diameter applies for wire sizes from less than 0·1 mm. to more than 1 mm., and holds throughout the operating range of temperatures.

The life of a tungsten filament is therefore directly proportional to its diameter for the same temperature, but as the heating current varies as the 3/2 power of the diameter, economic and other considerations (e.g. end-cooling and magnetron effect) preclude very long lives being obtained simply by using large diameter wires.

Knowing that a filament will burn out when evaporation reduces its diameter to 90 % of its original value, we can readily and accurately calculate its life. Early life estimations for cooled-anode valve filaments using tungsten evaporation figures of Jones and Langmuir.† gave a theoretical life of about one-half the observed average. More recent simultaneous determinations of the emission and rate of evaporation of tungsten by Reimann, using samples of tungsten from the same source as our filaments, show Jones and Langmuir's figures to be high by a factor of about 2, and the estimated life using Reimann's results is in excellent agreement with the actual lives obtained.*

Glass valves in general give lives of about one-half the theoretical. This is probably the result of a variety of causes, the most important of which is sputtering of the filament due to the fact that the vacuum in glass valves is less perfect than in water-cooled valves. Another contributory factor is the operation of batches of glass valves at a common filament voltage, the exact voltage being decided by the worst emitter, as distinct from the individual adjustment to marked volts for cooled-anode valves.

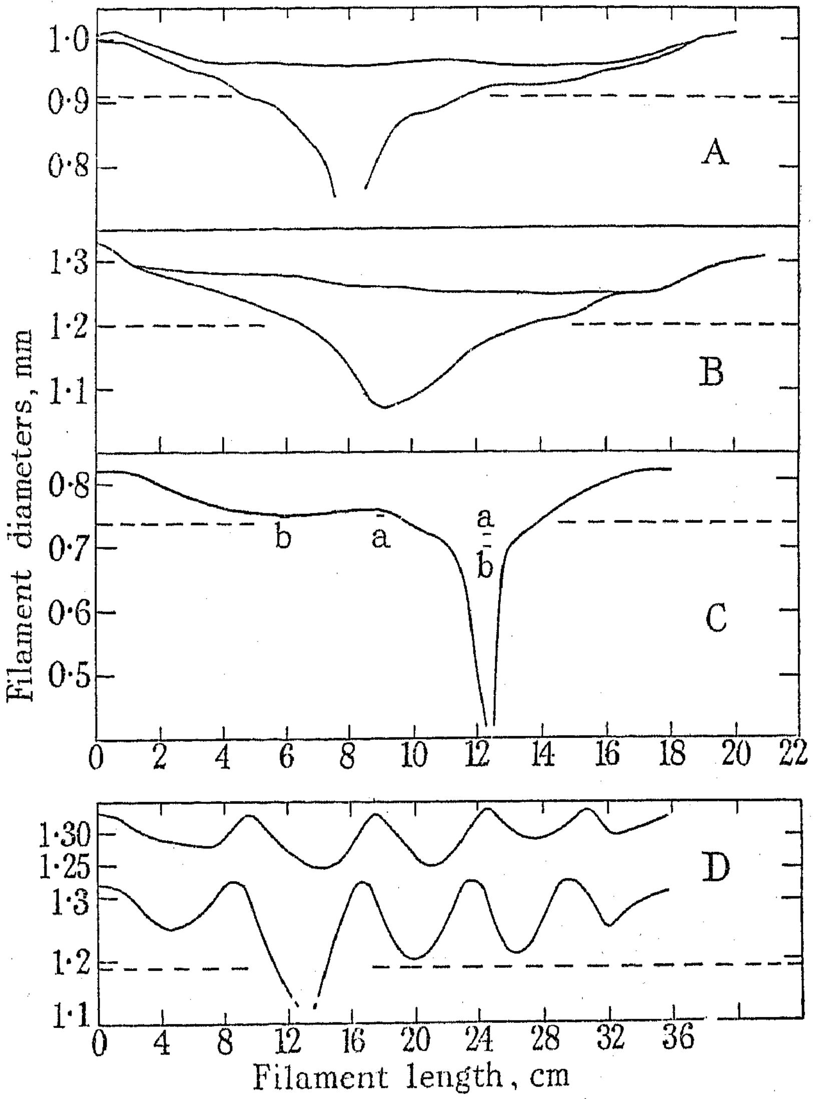

Fig. 17.—Reduction in filament diameter through wastage in operation.

A. Valve type C.A.R.2; 6 335 hours (overburnt about 5 %).
Both limbs shown, one only burnt out.
B. Valve type C.A.R.4 (V.T.36); 19 400 hours.
Both limbs shown, one not quite burnt out.
C. Valve type C.A.T.6; 10 817 hours.
One limb of one loop shown in full.
a. Measurements at points on companion limb of this loop.
b. Measurements at points on companion loop, wastages on both limbs equal.
D. Valve type C.A.T.10, early specimen with too many bracers. One of 3 burnt-out limbs shown, and one of 3 companion limbs of other polarity.

Broken lines show levels for 10 % loss in diameter.

As already described, it is the general practice to operate the filaments of cooled-anode valves at a temperature just sufficiently high to satisfy the emission demands of the circuit. Filaments are therefore rated on an emission basis, individual valves being stamped with the filament voltage required to produce a certain emission (marked voltage). A 5 % increase in this voltage increases the emission by 50 %, but reduces the life to one-half. This rapid variation of life with filament temperature necessitates the use of sub-standard instruments for the adjustment of filament voltage if the maximum life is to be obtained.

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* See Reference (15).

† *Ibid.*, (5).

* See References (5) and (16).

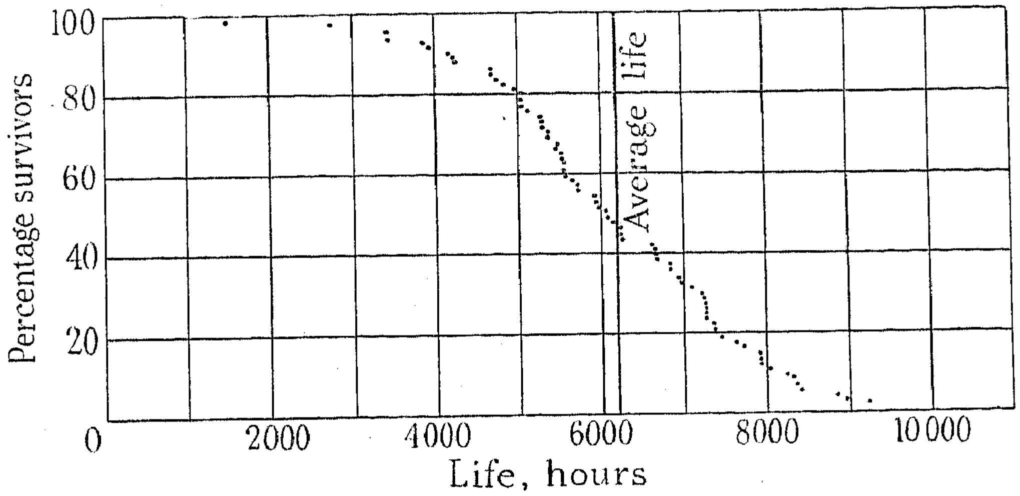

Fig. 18.—Survivor curve for 74 valves, type C.A.T.6, operated at "marked volts" in B.B.C. transmitters.
Failures for filament burn-out only.

Fig. 18 shows a survivor curve for 74 C.A.T.6 valves operated at marked volts in B.B.C. Regional transmitters. This curve is for filament burn-out failures only, and is typical of cooled-anode valves operated at marked volts. Some 60 % of the failures occur within ± 20 % of the average life, a few per cent of the valves fail at half the average life, and a few per cent survive to almost twice the average. The average life agrees very closely with the calculated value.

It may be asked whether multiple-filament valves last as long as the equivalent number of smaller valves. The answer is that apparently they do. This would only be so if the behaviour of all the limbs were very nearly equal. For although the parallel connection of unequal limbs must have some stabilizing effect in that initially thicker limbs will run hotter until the consequent more rapid evaporation has produced equality of diameter, yet initial equality is necessary since one-half of the limbs is in series with the other half. The limbs of each system are therefore carefully matched for diameter, and also all the wires are drawn with the same smooth compact surface, to equalize their thermal radiation properties. This last point is, as we have seen, of great importance.

It has been found experimentally that for valves operated at constant filament voltage the emission falls gradually during life. This is to be expected, since at constant voltage, apart from any change in surface condition, the temperature falls slightly during life, and of course there is the reduction in emitting area due to wastage. Emission measurements on long-life valves have shown that the drop in emission is rather greater than would be expected from the above considerations. The difference is due to the fact that the filament becomes less brightly polished during life, with a consequent further reduction in temperature and emission. If the surface condition of the filament remained uniform constant emission during life could be approached by putting a resistance equal to one-third that of the active filament in series and maintaining the whole at constant voltage. The resistance of the leads and of the cool ends therefore gives some compensation.

Constant filament-current operation cannot be tolerated, as in that case temperature and emission increase rapidly, and the life is only about one-fifth of the constant-voltage life. Constant emission is the ideal method of operation, and so far this has only been attained by increasing the filament voltage during life according to a predetermined schedule based on the results of emission tests on long-life valves. The survivor curve obtained by such a method of operation is shown in Fig. 19, which is for a batch of C.A.T.6 valves operated in the London Regional transmitter. Comparison with Fig. 18 will show that the curve is much steeper, i.e. the spread is much reduced, and the extra-long-life valves are eliminated. The extreme sharpness of the shoulder is

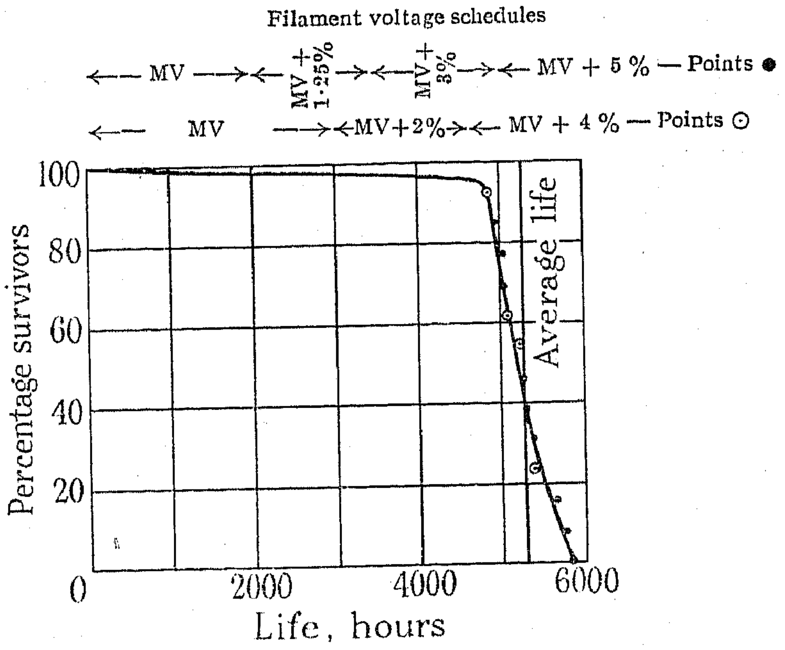

Fig. 19.—Survivor curve for 13 valves, type C.A.T.6, operated under conditions approaching constancy of emission during life in the B.B.C. London Regional transmitter.

The filament voltage schedules indicated are approximate.
Failures for filament burn-out only.
MV = Marked volts.

partly accounted for by the fact that an increment of filament voltage comes just before this point. The smallness of the overall spread is a verification of the accuracy of the original emission test and its extrapolation.

The average life under these conditions is reduced by about 15 %, but even so the practice of constant emission operation will tend to grow, since with constant voltage the filament must be over-run originally if it is to have sufficient emission towards the end of its life.

Table 1

VALVE TYPE, C.A.T.14.

Serial No.	Life, hours	S = Survivor F = Failure	Cause of failure.
Record of all C.A.T.14 valves used at the Beromünster Transmitter up to 31st December, 1937.			
13	8 049	S	—
14	4 724	S	—
15	8 014	S	—
16	4 701	S	—
Record of all C.A.T.14 valves used at the B.B.C. Droitwich Transmitter up to 31st December, 1937.			
1	7 228	S	—
2	3 127	F	Partly F.B.O.
3	9 483	F	F.B.O.
4	7 660	S	—
5	12 443	F	Leaky bellows*
8	1 421	F	Leak in foot tube*
9	1 896	F	Leaky bellows*
9A	5 162	S	—
10	5 569	S	—
11	8 985	F	F.B.O.
12	1 519	S	—
25	254	F	Leak in anode soldering
26	2 260	S	—
27	1 749	S	—
29	1 864	S	—
46	2 017	S	—
64	1 312	S	—
68	255	S	—
69	211	S	—
Record of all C.A.T.14 valves used at the Motala Transmitter up to 31st December, 1937			
17	10 885	S	—
18	7 249	F	Blue glow
19	4 162	S	—
20	10 004	S	—
21	10 743	S	—
22	1 739	S	—
23	190	S	—

* These valves were manufactured before the trouble with leaky bellows had had time to develop in C.A.T.10 type valves at Warsaw and Rugby.

Filament life can be considerably reduced by over-rapid switching, but such failures can be distinguished from normal failures by the nature of the break. Normal failures show a gradual taper on either side of the actual burn-out, while for filaments which have been over-switched the open circuit is more in the nature of a break than a burn-out, and in severe cases the filament may be distorted through the magnetic forces.

The absence of locally-accelerated evaporation in such cases seems to show that the real cause of the instability may be a local increase in resistivity, due to some internal change, rather than an increase in resistance due to reduction in diameter. It is only when the onset of instability is slower, as in the normal case, that there is time for the usual tapering towards the point of failure to develop. That the cause of instability is not the general loss of diameter according to the 10 % rule is also shown by the fact that, if the metal is removed rapidly by sputtering in a poor vacuum, the filament remains stable even when the diameter has been reduced by much more than 10 %. However, in valves as we know them the 10 % rule is still the expression of a reliable symptom of distress, even though it does not directly represent the actual disease.

Table 2

VALVE TYPES MENTIONED IN THE PAPER

Type	Filament			Max. Anode* voltage, d.c.	Max. Anode dissipation, kW	Normal wave-length (L=long S=short)
	Volts (nominal)	Amperes (nominal)	Emission (90 % saturated) amperes			
High μ Triodes						
VT26	22	41·5	5·5	12 000	10	L
CAT6	18/20	75	10	12 000	12	L
CAT10	30	220	35	15 000	50	L
CAT12	30	220	35	18 000	75	L
CAT14	32·5	460	100	20 000	150	L
CAT17	32·5	460	100	15 000	150	S
CAT19 (E8 30)	31·5	660	150	20 000	150	L
Low μ Triodes						
CAM3	17	70	3/7	12 000	12	L
CAM4	20	75	10	15 000	16	L
CAM5	28	325	22·5	15 000	60	L
CAM6	20	190	14	15 000	30	L
CAT12A	30	220	35	18 000	75	L
Rectifiers						
CAR2	18/20	50	4	15 000	5	
CAR4	18/20	75	7	15 000	6	
CAR6	18/20	120	10/20	20 000	20	

* Subject to certain restrictions as to circuit protection against flash-arcs.

The great majority of valves fail through filament burn-out, and only a small percentage through other causes. The effect of other failures on the survivor curve is to flatten the top bend, and hence to reduce the average life. For an established type, the percentage of other failures is very small, and the overall average life approaches the theoretical filament life.

For the largest valves, operation has not yet proceeded far enough to provide data for a survivor curve, but Table 1 shows the complete record of the three stations in which these valves have been in use longest.

Inspection of the serial numbers of the valves suffices to show that the occurrence at Droitwich of the majority of the failures for causes other than filament burn-outs is due to the fact that that transmitter was the pioneer in the adoption of the C.A.T.14 type.

In conclusion, the authors desire to tender their acknowledgments to Marconi's Wireless Telegraph Co., Ltd., and the General Electric Co., Ltd., on whose behalf the work described was done. They also wish to thank the Post Office, the British Broadcasting Corporation, and other operating authorities, and especially the members of the staffs of the various transmitting stations, whose reports of valve performance, compiled in great detail and at frequent intervals, have been most helpful and encouraging throughout the development of high-power valves.

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DISCUSSION BEFORE THE WIRELESS SECTION, 30TH MARCH, 1938

Mr. H. S. Walker: The early cooled-anode valves frequently went "soft" after a few hundred hours of operation; and therefore in 1930 the B.B.C. fitted in their regional transmitters a two-range microammeter which could be used for the measurement of grid current when the transmitter was operating, or, when the shunt was disconnected, for measuring the backlash current. The gas tests described by the authors were used originally by us to detect when valves were going soft, so that they could be removed before they caused an interruption in the service. Nowadays, however, as the vacuum remains nearly constant, there is little use for this test. Nevertheless, since 1930 the engineering staff of the B.B.C. have taken 38 500 gas-test readings.

These readings, particularly the ones on C.A.T.14 valves at Droitwich, have revealed a new phenomenon. Shortly after Droitwich was opened in June, 1934, one of the C.A.T.14 valves, No. 9*a*, showed a backlash current of about 400 microamperes as compared with the value of 150 microamperes at the normal anode current of about 2 amperes. The organization which the authors represent were informed of this, and proceeded to repump the valve. When it was returned to service a normal backlash figure was obtained. Some months later, however, the backlash figure began again to increase, and after about another 1 000 hours a figure of 750 microamperes was obtained. The valve was examined for leakage but none was apparent, and it was again repumped. After about another 1 000 hours the backlash increased fairly suddenly to the rather large figure of 2 500 microamperes, and finally to 5 000 microamperes. So great was this figure that, if it had been due to softness, the valve would have shown a blue glow, which it did not. The valve in fact worked perfectly normally and was therefore kept in circuit; after a week the backlash had fallen to some 400 microamperes, at which figure it has remained. Since that time the valve has been in operation for about another 2 000 hours.

Recently another valve, No. 83, has shown similar symptoms, and, as they could be reproduced without application of the anode voltage, a leakage in the circuit or across the valve seemed to be a possible cause; but practical tests did not confirm this suggestion. I think that the cause of this apparent backlash current must be some form of primary emission from the grid, due to heating of the grid or radiation from the filament. Since pure tungsten has an affinity for thorium, it is possible that during pumping operations a small quantity of thorium or sodium was taken up by the grid. Various causes, such as sudden changes of temperature at the grid, might result in the thorium diffusing to the surface, and there producing an emission which might account for the phenomenon I have described. Such irregularities of emission are not unknown in valves fitted with thoriated-tungsten filaments. It is reasonable to suppose that, if this explanation is the correct one, the two repumping operations which the authors' organization carried out were unnecessary. I should be interested to have their observations on this point.

Reliability of operation and freedom from breakage are more important in broadcast transmissions, at least as far as the B.B.C. is concerned, than in any other service in which valves are used. At the same time the distortion factor of a transmitter must be kept down to a very small value: a figure of about 4 % at 80 % modulation is the most usual. This demands a much more nearly linear characteristic in the valve than is required for telegraphic or commercial telephonic use, and, since this must be maintained through life, a schedule of gradually increasing filament voltage is applied to B.B.C. stations. Fig. 19 shows one of the schedules adopted, which applies to C.A.T.6 valves in B.B.C. Regional transmitters. We have found by practice that 1·15 amperes per valve is the maximum we can work at and maintain linearity with reasonable economy of filament, and it is on this basis that the schedule of increasing voltages shown in Fig. 19 has been built up. Under these conditions C.A.T.6 valves in B.B.C. Regional transmitters averaged 5 816 hours during 1937, taking all failures into account. (The slightly higher figure quoted by the authors includes the filament burn-outs only.) If it were not for the stringent conditions of linearity demanded by our service, very much better figures could be obtained. We have heard of foreign stations with C.A.T.6 valves which have lasted for 10 000 hours, but the distortion must have been greater since the filaments were run at a lower temperature.

The authors state that the electrical conductivity of cooling water should not exceed 370 reciprocal megohms per cm. cube. While such water is doubtless quite safe from the valve point of view, it would give rise to all sorts of troubles in other parts of the equipment, particularly electrolysis down the water hose columns. A maximum safe figure is much more likely to be of the order of 100 reciprocal megohms per cm. cube. The water used by the B.B.C. is much purer, with a conductivity of the order of 30 reciprocal megohms per cm. cube.

Mr. S. R. B. Pennington: As indicated in the paper, we are considering adopting forced-air cooling for the external filament leads of new high-power cooled-anode valves, instead of water-cooling. This change should simplify the cooling circuits, but it may give a lower factor of safety.

This important question of cooling-circuit simplification has led to a consideration of the use of air-cooling for the anodes of larger valves. The largest size to which air-cooling has so far been applied has an anode-dissipation rating of about 5 kW and requires some 300 cu. ft. of air per minute for cooling purposes. These are not yet in commercial use, but an air-cooled version of the ultra-short-wave valve C.A.T.15 has been in service in the B.B.C. mobile television transmitter since May, 1937.

The authors refer to the difficulty of obtaining early life data for the larger valves: in this connection we have been fortunate in having the collaboration of the B.B.C., in that their progressive policy has enabled us to supply them with apparatus embodying many of the larger-type valves immediately they have reached the production stage. Examples are the C.A.T.14 valves at Droitwich, and the short-wave version of this type in the high-power short-wave transmitter recently installed at Daventry.

The short-wave C.A.T.14 valve has a separate circular grid seal, an anode dissipation of about 150 kW, and a filament emission of about 100 amperes. Although much work has been done, we have not yet determined the limiting wavelength at which the performance of this valve can still be considered satisfactory, i.e. the wavelength at which transit-time effects and the resultant grid-load call a halt. At 13 metres, however, the anode conversion efficiency still holds approximately to the long-wave value, although there is some indication of losses introduced by electron transit-time effects as the grid input has risen slightly above the general level at the longer wavelengths. With this valve operating in a 13-metre circuit as a modulated Class C amplifier at the normal input, a carrier output of 50 kW at an anode efficiency of about 70 % is normal.

Finally, in their schedule of operating conditions the authors mention the precautions essential to good valve life. To these I would like to add the suggestion that station authorities should see that all spare valves are periodically tested under normal working conditions, so that damage or defects occurring during storage may be detected at an early date and a known good stock held.

Mr. B. N. MacLarty: I am glad to learn that it may be possible in the near future to adopt air-cooling for the anodes of larger valves. Such a practice is more feasible now than it was a few years ago, owing to the introduction of high-efficiency systems of modulation in which the valves are worked at much less than their rated anode dissipation. I can, however, visualize some difficulties in connection with the introduction of the air to the valve and its removal. Will it ever be possible to air-cool a valve of the C.A.T.14 size? I am particularly concerned regarding this matter, because the cost of the cooling system in a high-power transmitter is a considerable part of the total cost of the station, and I think it is now advisable to consider either working at higher water temperatures or adopting some other method of cooling.

The question of the water cooling of filament and grid seals is also very important. Owing to the small volume and low velocity of the water passing through these seals we get trouble due to the settlement of fine sediment inside the cooling labyrinth, and it will be an improvement if future high-power valves have their seals cooled by air.

The possible trouble mentioned by Mr. Pennington, regarding the failure of the air supply in an air-cooling system, can be overcome by adopting a principle similar to that which we have adopted on water-cooling systems, namely two pumps in parallel either of which is capable of dealing with the total load.

The fact that the magnetic field due to the filament-heating current can sensibly modulate grid current, is a matter of interest, especially as we are nowadays being pressed to adopt alternating current for heating filaments. The authors' comments on this matter indicate that great care is necessary in the design of valves in which alternating current is to be used for heating the filaments.

Although valves are in use having 3-phase filaments which give remarkably low noise-levels, it appears that their success in this respect may be to some extent fortuitous. The authors mention that valves with single-phase filaments connected by Scott transformers give a lower hum level than 3-phase valves. This may apply to the particular valves tested by them, but I am not convinced that the subject has been sufficiently investigated. Are the authors aware that a remarkably low hum level has been obtained from a single C.A.M.5 valve, working as a modulator and heated by 50-cycle alternating current?

Mr. L. Grinstead: I should like to know what the authors' experiences are with the use of tungsten for grid-building purposes, as I have found in general that tungsten grids are much more prone to secondary emis-

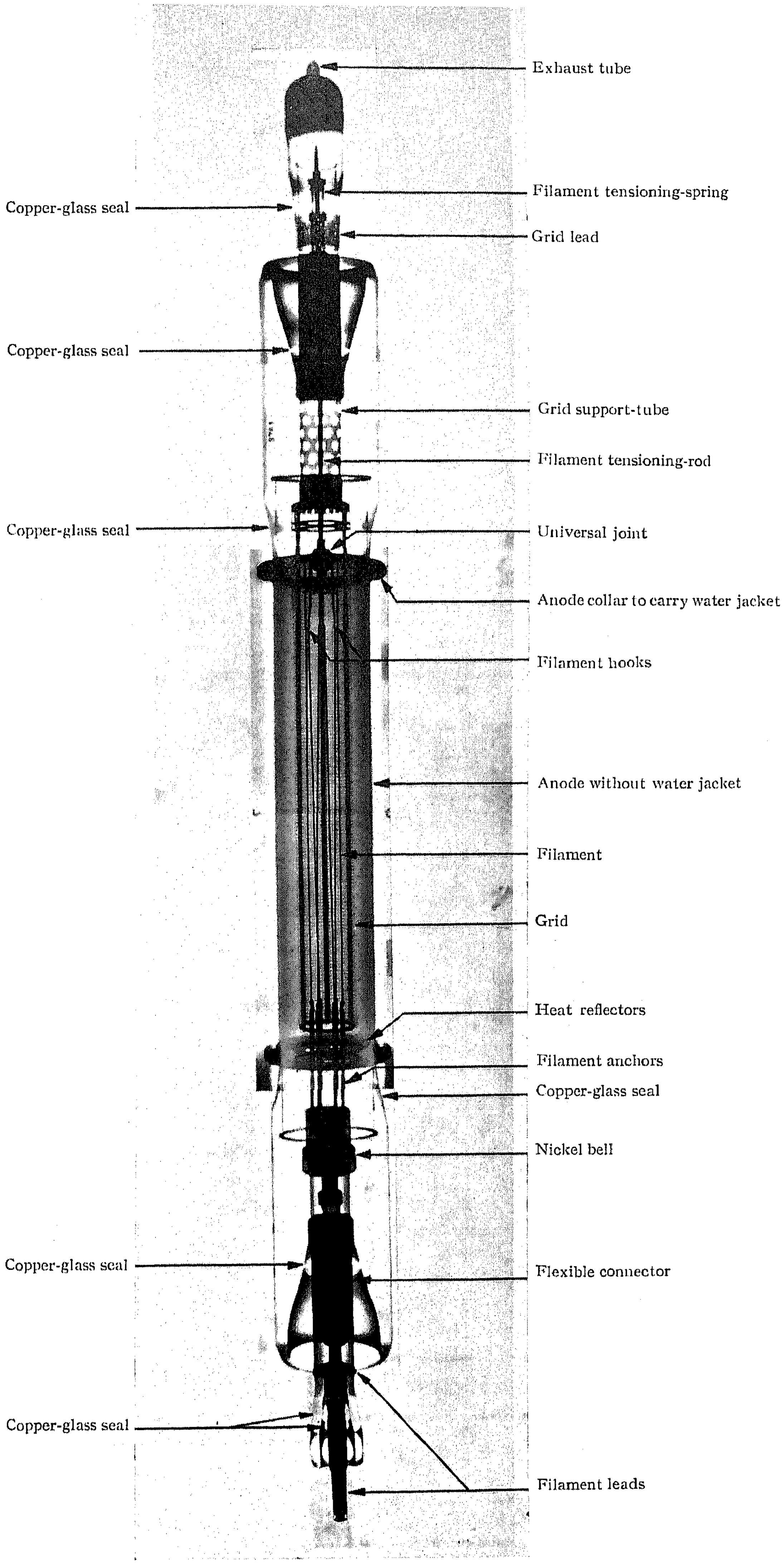

Fig. A (see Mr. W. T. Gibson's remarks).

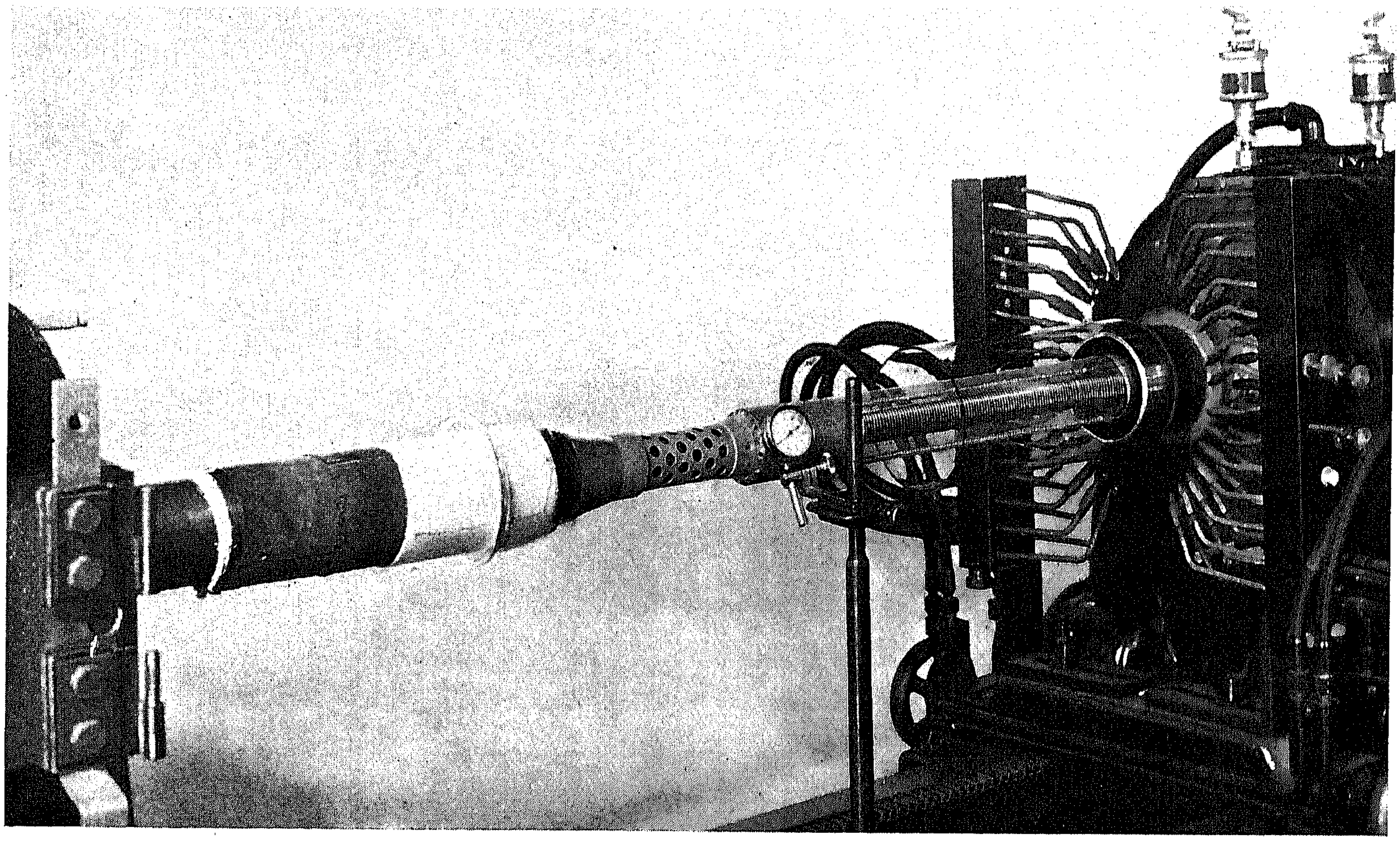

Fig. B (see Mr. W. T. Gibson's remarks).

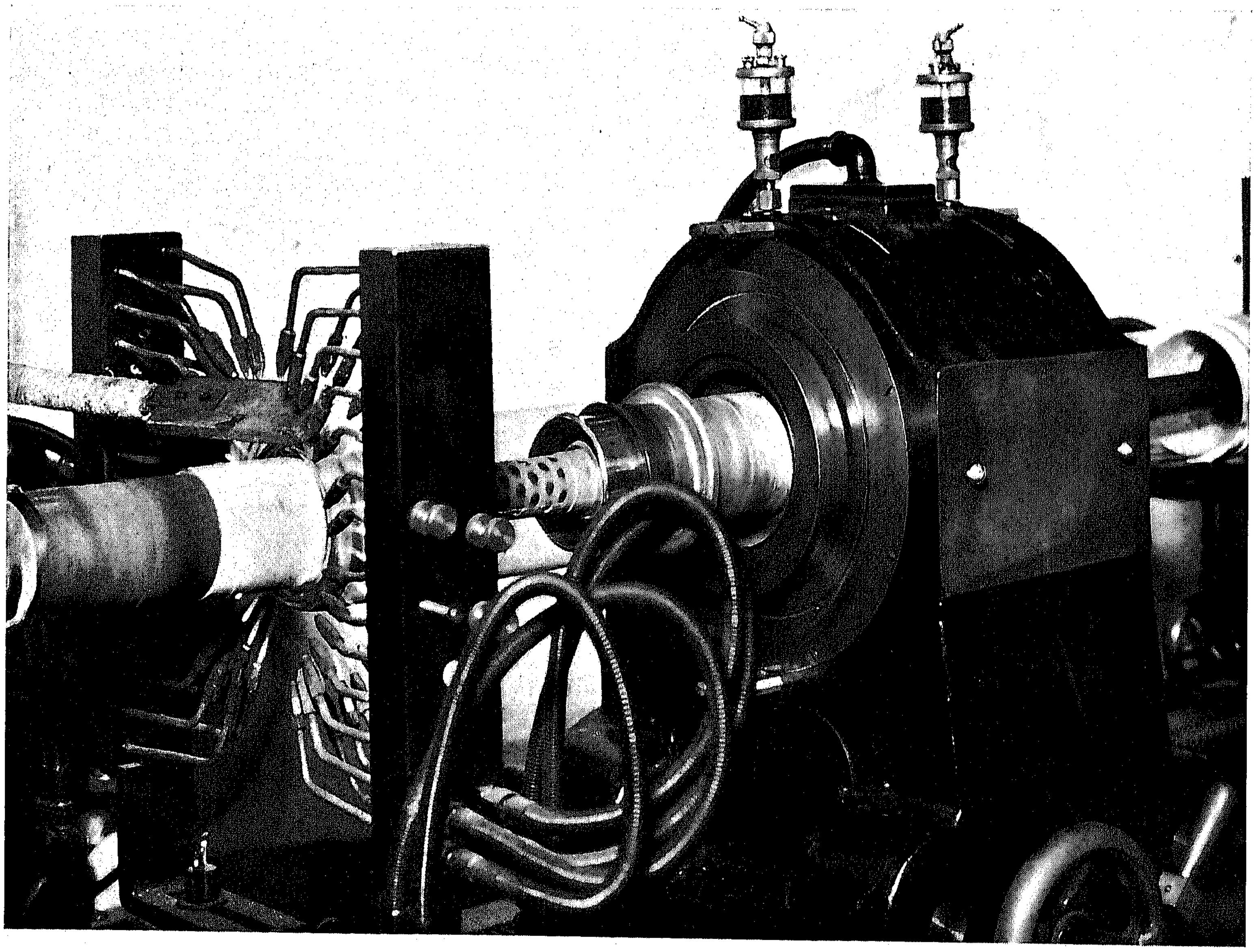

Fig. C (see Mr. W. T. Gibson's remarks).

sion than are molybdenum grids. An interesting point is referred to on page 184 in connection with the photo-electric properties of molybdenum and tungsten. Is it possible that these photo-electric effects have some bearing on the increased secondary emission observed from tungsten?

The authors speak of the possibility of employing a grid lead-out in the form of a metal cylinder sealed between two sections of the glass envelope; this no doubt makes a very good attachment to the grid itself, but would it be so easy to make a physical connection from a high-frequency circuit to such a cylinder? I believe that a grid seal of this description would not show much advantage over two well-constructed side seals of more or less conventional type.

As regards this question of seals for grid and filament connections, valves have been made for some years with chrome-iron seals carrying quite high values of radio-frequency current, but usually these seals are made in the form of thin discs. No difficulty has been experienced with them, although the authors appear to have had unfavourable experience with chrome-iron seals.

In connection with Figs. 7(*a*) and 7(*b*), I notice that actual tests are made for values of emission up to approximately 10 % of full rating and that the authors have also plotted on these diagrams one or two points as additional observations for checking the extrapolation. Apparently a diode is used for the check values, and while this method disposes of the problem of power dissipation it does not eliminate the effect of space current on the distribution of filament temperature. Are these check observations taken by the intermittent method referred to early in the paper?

Finally, I should like to ask for a little more information concerning the authors' statement that the practice of constant-emission operation will tend to grow. It seems to me that while in theory it is desirable to have valves running under constant-emission conditions, fundamentally it is not of great importance provided there is enough emission initially to meet circuit requirements and good valve lives are obtained. In the results shown in Figs. 18 and 19 it appears that substantially constant-emission operation is obtained at the expense of 15 % of average life, which I think is rather uneconomical from the user's point of view.

Mr. L. B. Turner: The development of the high-power valve marks, as far as we can guess, the end of the changes in the method of producing high-power high-frequency current. We have seen developed and discarded the spark transmitter giving damped trains, the rapid-spark modification giving an approximation to continuous-wave working, the Poulsen arc, and the various patterns of electromechanical alternators with and without subsequent frequency multiplication. All these were shown to be capable of producing an output of some hundreds of kilowatts; but all were relatively complicated and inflexible contrivances, and they have been set aside in favour of the monument of simplicity shown in Fig. 1. After explaining their empirical procedure, the authors say that "the further it is removed from the plane of trial and error by increasing insight into what is actually happening in the valve in manufacture and in operation, the more quickly will improvements be made and faults eliminated." A most satisfying feature of their work seems to me the thoroughness with which this principle has been acted upon.

Many interesting empirical discoveries and theoretical investigations were made by the authors: I want just to refer to one of each. First, who would have anticipated that the problems of water-cooling of the copper anode would prove to be seriously influenced by the streakiness of the electron bombardment owing to the focusing action of the grid wires (as described on page 182)! Second, in seeking to obtain uniformity of filament emission in different valves of a type, how elegant is the statistical analysis of Fig. 8, where a diagnosis is got from the axes of scatter in the target diagram of volts plotted against amperes for equal emission!

In triodes for transmission, as in those for reception, a long extent of sensible straightness of the characteristic is often desired: but, as far as I have been able to ascertain, the straightness obtained has been determined by accident rather than design. In Fig. 10, it seems that from C to B (and I do not understand why not also below C) the curved foot of the $\frac{3}{2}$-power space-charge law is modified by the magnetron cut-off effect; and I think this effect must be a straightening one, since it becomes less apparent as the anode potential rises. From B to A we have the $\frac{3}{2}$-power law, but over a limited range which is itself fairly straight; and above A, the continued upward concavity of the $\frac{3}{2}$-power law is countered by incipient saturation of the non-uniform filament, tempered by the telescoped Schottky effect described on page 190. As the happy result of all those actions the characteristic is said to be straight from C to D—say from 1A to 4A when the saturation current is 5A. Again I venture to ask; are the effects thus analysed consciously cultivated and balanced as part of the design?

With reference to Fig. 11 it is explained how the full-emission characteristic may be calculated from the curve observed with an under-heated filament. Although I think the theoretical basis of this procedure must be right, I cannot follow the argument on page 192. I suspect there is a logical hiatus; for the validity of the construction must surely depend on the log/log plotting in Fig. 11, and this is not referred to in the argument on page 192.

Four brief points in conclusion: (i) Will the authors say what, if any, is the distinction (on page 183) between "equilibrium pressure" and the actual pressure in the valve? (ii) On page 186 "heating current H leaving the positive end of the filament" is, I think, a misprint for "entering the positive end"; and, if so, the words "local temperature" four lines lower should apparently read "temperature at the negative end." (iii) Can Fig. 1 be supplemented by a statement showing the linear dimensions of the C.A.T.14 valve and giving some rough indication of the cost? (iv) What happens to the 2–3 oz. of tungsten evaporated during the life of a large valve? Is it ionized by electron collision? Is it deposited on the grid or on the anode?

Mr. W. T. Gibson: I should like to give a short account of a valve made by the company with which I am associated, which shows how similar problems to those dealt with by the authors have been solved by rather different methods.

Some time ago we had the task of developing a valve, Type 4030, which was to have a useful emission of 45 amperes and to be suitable for operation on long or short waves. In view of past experience, it was decided to make it double-ended, and it was known that it would probably be required to operate with the filament terminals at the lower end. As a result, what we believe is a unique type of construction was adopted (see Fig. A, Plate 3, facing page 200).

The filament leads consist of two concentric copper tubes, joined to the filament stem by copper-to-glass seals, and there is a flexible connection made of copper cable in the inner lead to allow for relative expansion. The two filament leads are joined positively together inside the valve by means of a steatite insulator which separates them electrically. This insulator is located inside, and completely surrounded by, a highly polished nickel bell from which the six filament anchors project.

The filament consists of three hairpins of tungsten without any central support. The bights of the filaments are carried by hooks which are mounted on a ball-and-socket universal joint consisting of a tungsten ball engaging in a molybdenum socket carried by a molybdenum rod which passes through the grid stem, which is again a copper tube sealed to the glass. The rod passes through two silica insulators which occupy a cold portion of the tube, and is then pulled by a spring which is adjusted very accurately to the required tension.

The development of this valve was very greatly helped by the use of a glass-working lathe which we constructed some years ago. This lathe has a bed about 12 ft. long, three heads with special chucks, and two fire heads. The machine is of the very highest precision, equal to that of a high-class metal-working lathe.

As the valve has an axis of symmetry, it is possible to take full advantage of the precision of the lathe, and as an indication it may be stated that the grid is checked for truth with a dial gauge on the lathe immediately before being sealed in. The lining-up of the grid is shown in Fig. B (see Plate 4), and the sealing-in in Fig. C (see Plate 4). The filament is sealed into the opposite end, a tool being inserted through the centre of the filament stem to hold the assembly at this stage. The lathe is provided with an electric annealing oven which allows the complete annealing of the glass without disturbance of the geometrical relationships. When the valve has been removed from the lathe, the tool which holds the filament structure is withdrawn and the end of the filament lead closed by brazing-in a copper disc.

After the valve has been exhausted, the water jacket is soldered on. This water jacket is so constructed that the cold water enters at the middle and flows beneath the baffles, giving a thin film flowing at high velocity over the heated surfaces. The water is then collected from the two ends and taken to the outlet pipe. The water jacket is corrugated, to provide flexibility to compensate for the relative expansion of anode and jacket. The valve operates silently and with perfect safety on test up to 160 kW dead loss.

The precision obtainable by these methods has made it possible to produce valves of extremely high constancy and regularity of characteristics. For example, in a recent group under normal manufacturing conditions, 85 % of the valves had an anode current within ± 4 % of the rated value, and 85 % have filament volts and current within ± 1·5 % of the mean.

The connections to these valves are made through cast aluminium radiators which are jointed to the filament and grid leads. Ample cooling is provided by the natural circulation of air, forced air or other cooling being unnecessary. In order to give an idea of the factor of safety, I may say that during the pumping of the valve the anode is operated at 800° C. and no radiators are used on these electrodes. Connections of very small section are made to the electrodes, and no auxiliary cooling is employed even under these circumstances.

The importance of the contribution of photo-electric current to total gas current mentioned by the authors is clearly demonstrated in the testing of this type of valve, and Fig. D shows the frequency of occurrence of any given value of gas current among a large group of valves. It will be seen that the maximum frequency occurs at the minimum recorded value.

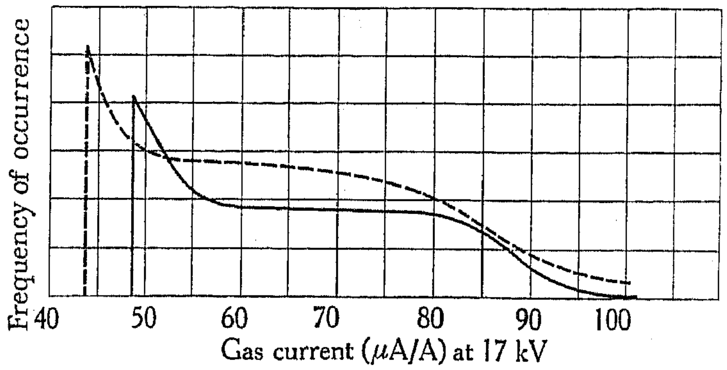

Fig. D

The method which we use for estimating the total emission is rather different from that described by the authors. We use Davisson power-emission paper, and make measurements of anode current up to about 10 amperes. On this paper the total-emission lines are absolutely straight, and the ratio between the filament watts at the required emission and at, say, 10 amperes emission is constant to within less than ± 1 % from valve to valve.

These valves are in use in a number of large stations, including the Daventry Empire Stations, and the following record of valve life of all valves in the Liblice station indicates the satisfactory nature of the performance:—

Valve serial Nos. 6032, 6192, 6031, 6054, 6006—7 854 hr. 15 min.; No. 6042—4 747 hr. 0 min.; No. 6140—3 107 hr. 15 min. All except the last-named were survivors.

Mr. J. Greig: I regret that, presumably on account of space limitations, it has not been possible to treat more fully the inter-relations between valve characteristics, circuit conditions, and linearity requirements.

I should like to comment on one or two points relating to the design of valves of higher powers for operation on short wavelengths.

The provision of multiple seals or ring seals for the carrying of heavy grid or screen currents, and the difficulty of obtaining uniform distribution of such currents, are referred to in the paper. This non-uniform distribution, resulting from the different effective inductances of the various current-paths within the valve, in relation to the disposition of the external circuit, would appear to be to a large extent inevitable, and where the current is conveyed to the active portion of the grid by the grid supports themselves, some distortion of the structure due to unequal heating is to be expected. It would be interesting to know whether it might be considered desirable and practicable to by-pass the high-frequency currents from the grid supports by relatively flexible strip conductors more suited to the carrying of high-frequency currents, making contact with the grid as near as possible to its active portion where the temperature would, for the most part, be determined by radiation from the cathode and by grid energy dissipation.

The necessity for reducing to a minimum the inductance of the path between the grid seal and the active portion of the grid is well recognized, but the most inherently inaccessible electrode of the valve is the incandescent cathode, the true " earth " of the circuit, and it seems probable that the inductance of the cathode and its associated leads may not have a negligible effect at short wavelengths. It is likely that some of the loss of efficiency in short-wave circuits, in the region before transit-time effects become significant, may be due to the inductance of the cathode circuit. Certainly the difficulty of anchoring the screen to cathode potential and of avoiding parasitic oscillation must be enhanced by this inductance. Perhaps the authors could say whether cathode inductance has so far been found to have any appreciable deleterious effect.

Mr. E. B. Moullin: With regard to the telescoped Schottky effect mentioned in this paper, I am looking forward to the publication of the results of an investigation on this subject by Mr. G. W. Warren. It will be interesting to know how this subject has been affected by Mr. E. W. B. Gill's recent paper,* which seems to show conclusively that the Schottky effect does not follow the Schottky law, but varies as the inverse half-power of the voltage.

It is amazing to learn that the life of valves is now so predictable a quantity and that it is so closely related to the diameter of the filament.

Messrs. M. Ponte and **R. Warnecke** (France) (*communicated*): Our experience indicates that insulators which are subjected to high voltages at high frequency in the hot parts of thermionic valves do not give satisfactory results.

We think, like the authors, that the practice of fitting insulators between the grid and the anode is impracticable because of the excessive voltage which normally exists between these electrodes. We have found, however, that insulators between grid and anode can be adopted even with valves of the highest power without introducing any difficulty from the point of view of operation; it is in fact employed in nearly all the types of high-power valves used by the Société Française Radio-électrique. Valves constructed in this way are very robust and show great uniformity in their characteristics. The presence of a central guide makes the fitting of a filament-grid insulator relatively easy. It permits of a solid grid-filament construction which cannot be mechanically distorted in the transverse sense, yet allows expansion of the grid, central guide, and cathode.

The usual shape of valves enables insulators to be employed in situations where the temperature is low, without congestion and without increasing the size of the valve. For example, in a valve having a normal carrier output of 30 kW with Class B amplification the anode is only 530 mm. high and 113 mm. in diameter.

As regards grid seals in the form of a ring, we would draw the authors' attention to the difficulty of degassing these metal parts, which, it seems, must become heated in operation.

In reply to critics of the practice of fixing the grid by means of two arms sealed to the glass envelope, this system has a rather important advantage, especially for valves operating on short waves. By connecting the circuit formed by the grid and its two supports to a source of current it is possible to heat the metal of the grid sufficiently to degas it up to the level to which it is possible to heat the grid by electronic bombardment during the pumping process. This point is especially important in connection with short-wave valves, where large high-frequency grid currents are employed in normal operation. In valves rated at 150 A for normal use it is easily possible to pass a current of 1 500 A, i.e. to apply an overload such that the temperature-rise during the pumping process is much higher than that which exists in normal operation. This is a useful feature from the point of view of the final gas-absorption capacity of the sealed valve.

On page 179 the authors suggest that in valves having a low value of μ the grid may consist of rods parallel to the axis of the cathode. As no precise information is available as to the best value of μ to adopt for a valve operating as a Class B high-frequency amplifier, for example, it seems clear that when the authors refer to a valve with a low μ value they mean one which operates throughout the cycle with a negative grid voltage. This ought to be precisely stated, because it is impossible to imagine a valve with the grid made of rods parallel to the filament developing a positive grid-potential, due to the high grid-current of such a valve.

We do not agree with the authors that seals made of thick metal rods in glass may prove fragile. Our experience has shown, for example, that seals made of 10–20-mm. molybdenum rod in glass can be produced commercially without risk of breakage provided glass having the right properties is employed and a suitable sealing process is adopted. With regard to the use of seals on copper bezels, which we have employed for several years for filament-current connections, we find that direct seals made of rods have the advantage of fewer breakages and much greater ease of manufacture. By having a continuous conductor to carry the current one avoids troubles due to bad contacts or faulty bonding, both inside and outside the valve. This conclusion is supported by our experience as manufacturers and users of high-power transmitting valves. For copper-glass seals we have found it preferable to seal on both sides of the bezel with glass of a suitable coefficient of expansion.

We should like to know whether the authors have actually obtained all the expected advantage from anodes grooved on the outside from the point of view of the transfer of heat to the cooling water. We find that, for reasons not yet understood, increasing the surface area of the anode in contact with the water gives no marked advantage. When in use on transmitters, however, grooved anodes are subject to the practical difficulty that they "fur" more rapidly than smooth anodes. The interior of the grooves becomes coated with talc which has the effect of reducing the cooling area as compared with a smooth anode.

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* *Philosophical Magazine*, 1937, vol. 24, p. 1093.

The authors mention the possibility of blackening the interior of the anode in order to diminish the amount of heat reflected towards the envelope. A better way of achieving the same result is to employ channelling of suitable profile on the inside of the anode; this method seems preferable because the low stability of the deposit in the presence of heat sometimes limits the efficiency of degassing treatment, and also because the limited adhesiveness of the deposit may be the cause of flash-arcs.

The usual system of construction of valves does not always permit normal degassing of all the mechanical parts. Often the electrode supports and the current leads are only raised to moderate temperatures during the pumping process. Examination of data on the gas-emission of solids (particularly metals) seems to show that certain parts of the valve will give off gas constantly throughout its life because of the impossibility of raising them to a sufficiently high temperature during the degassing process. This gas tends to be re-absorbed by those parts of the valve which have been well degassed. The construction of the valve should therefore be such that the parts which cannot be heated during the pumping process are as few as possible. Our manufacturing experience shows that the degassing of very powerful valves is greatly helped if the heating of the anode is done by electronic bombardment instead of solely by radiation from the cathode; we think this to be due to the high activity (thermal or other) of high-speed electrons.

The application of a high voltage to the anode of a valve causes a fixation of the gas which reduces the pressure in the envelope; this fact may be confirmed by following the readings of a sensitive ionization gauge connected to the pumping tube of a large valve. These readings when plotted give a curve like that in the authors' Fig. 2. Was such an effect taken into consideration during the examination of their results?

The parallelism of the curves in Figs. 2 and 4, giving the backlash ratio and the photo-electric current, is striking; it was probably possible to investigate in these experiments the neighbouring points representing the K and L absorption limits of molybdenum. If the photo-electric effect is really due to the softest X-rays one ought to be able to find a curve which would give the law of variation of the continuous background emitted by copper with the accelerating voltage applied to the electrons, for the wavelengths in question. Have the authors made any investigations on these lines? Such investigations seem to us to be necessary because the authors rather under-estimate the importance of the primary emission from the grid, which most frequently sets the limit to the backlash current. This emission is due to the contamination of the grid by the particles detached from the anode under the influence of electronic bombardment. We may therefore consider that a state of equilibrium is reached at the surface of the grid which depends on the anode voltage and the grid temperature. The existence of backlash currents in the absence of an anode voltage, which vary according to the age and history of the valve, is evidence of this phenomenon.

We have examined the effect of the emission from the grid by methods enabling the grid to be heated by bombardment to normal working temperature and at the same time the backlash current of the valve to be measured in the absence of anode voltage. We find that the emission often begins at abnormally low temperatures and reaches high values under suitable working conditions. As, in certain instances, the phenomenon affects the value of the backlash current, even in the static condition which obtains during the process of measuring the space current, when the grid is only heated by radiation, we would ask the authors whether Curves A and B in Fig. 5 do not represent, partially at least, this heat-emission effect.

We have been able to verify, by different methods, the accuracy of the authors' hypotheses regarding the predetermination of valve characteristics. As regards the distribution of primary currents between the grid and the anode, the assumption that $I_g/(I_g + I_a)$ is a linear function of E_g/E_a is merely a first approximation which is very difficult to put to practical use. The terms which must be introduced into the second quantity in Lange's equation, namely

$$\frac{I_{g_1}}{I_{a_1}} = f\sqrt{\left(\frac{V_g}{V_a}\right)}$$

to take account of the space charge, do not reduce to a form justifying the linear relation mentioned above. We should like to know what the authors think of the use of the simple relation of Lange, by which it is possible either to calculate the distribution coefficient from the dimensions of the tube, making use of simple considerations derived from electron optics; or to measure the distribution coefficient at the points where $V_g = V_a$, the cathode being heated normally. We have found that for certain types of triodes, if corrections are made to eliminate errors due to the effect of the magnetic field and due to the secondary emission from the grid and the anode, Lange's law holds good as a first approximation when $V_g = V_a$ and the voltages are not too low; and it even holds for the case of an appreciable space-charge.

The reason why the greater part of the secondary emission from the grid comes from the ends of the grid is that, over the central portion, the primary space-charge retards the flow of electrons from the grid to the anode much more than that from the cathode to the anode. In this connection we would mention that merely by suitably choosing the position of the grid between the cathode and the anode it is possible to make the space charge between these electrodes act as a "stopper" grid. In valves of normal construction this phenomenon is related to the well-known effect produced by the magnitude of the total cathode emission on the distortion of the grid characteristic by secondary emission.

With regard to flash-arc breakdowns, some of our valves operate without troubles of this sort although their grid-anode distance is relatively small (20 mm.) and their working conditions are exacting (anode voltage

19 000–20 000 volts, with full excitation for a load which is constantly changing throughout the cycle of modulation). The fundamental factors which enter into the problem of flash-arcs seem to be the initial treatment of the electrodes and the nature of the impurities which may be deposited on their surface during the pumping process.

Messrs. J. Bell, J. W. Davies, and **B. S. Gossling** (*in reply*): In reply to Mr. Walker's points in connection with two C.A.T.14 valves at Droitwich, we feel that although they present similar phenomena the causes are different.

Valve No. 9a was in our opinion a straightforward case of softening, although at the time the cause was obscure. The possibility of a slow external leak proved, after extensive tests, to be out of the question. Grid emission was naturally a point which appeared to be an explanation for the high backlash, but after due consideration of actual facts there was no other alternative but to accept the high backlash as being due to high gas pressure.

When the valve was returned in the first instance, after some 1 000 hours' life, it was found that, out of some 700 μA reverse grid current, only 100 μA was due to grid emission. On the second occasion, after a further 1 000 hours' life, the grid-emission component of the total reverse grid current was even less, being only 25 μA in 1 000. From these facts there can be no doubt that the high backlash figures on this valve were due to softening.

During the history of cooled-anode valve development, we have, on various occasions, had reasons to believe that there do exist, in certain batches of metal, pockets or cavities which are not apparent on the surface. Such cavities would, under certain conditions, evolve gas at a slow rate and account for gradual softening. In more recent times we have some proof that such conditions are definitely possible. Cases have been examined where, on the inside surface of copper anodes, several blisters some few square centimetres in area and probably one half cubic centimetre in volume have appeared after heat treatment up to 500° C. In the early stages of gas escape from the pockets, the vacuum would be maintained by clean-up, but when the clean-up capability was exhausted the pressure in the valve would rise.

We do not think that the re-pumping of this valve was unnecessary. It was a clear case of softening, and, although at the time we were a little uncertain of the cause, more recent experience has proved that the action taken was justifiable.

The second (valve No. 83) referred to is, we agree, a case of grid emission, but we would emphasize that such cases are extremely rare in our experience. The suggestion that the presence of thorium would account for the effect is reasonable but for the fact that in accepting the theory one is faced with another question, i.e. the origin of the thorium. In all high-power valves the filaments are pure tungsten and grids molybdenum or a combination of pure tungsten and molybdenum. That the grid emission is due to an active substance on the surface of the grid cannot be disputed, but we are sure that the active matter is not thorium. Sodium, potassium, or even copper oxide, can be considered as possible causes, but to which of these any definite case of grid emission can be ascribed is very difficult to say. Unfortunately, when grid emission cases have been examined no positive conclusions have been reached, probably owing to the evidence being destroyed when air is admitted into the valve.

Mr. Walker gives a figure of 1·15 amperes as the mean anode current of a C.A.T.6 valve in B.B.C. Regional transmitters. If the distortion factor is to be kept down to the low value of 4 % at 80 % modulation, the total emission of 10 amperes must be available throughout life. A longer valve life can only be obtained either by sacrificing quality or by using a lower mean anode current whilst retaining the same low distortion factor. The average life of 5 816 hours which Mr. Walker obtains from C.A.T.6 valves confirms the statement in the paper that, for an established type, the average life approaches the theoretical filament life—in this case slightly over 6 000 hours—and shows that the operating conditions must have been in every way satisfactory.

Mr. Moullin will be interested in this check of the predicted life carried out under accurately controlled conditions.

In reply to Mr. Grinstead on the subject of constant-emission operation, the important point really is that if a valve, operated throughout its life at constant filament voltage, is to have enough emission at the end of its life not to increase the distortion factor, then it must have much more than enough emission initially. The average life under these conditions will be depressed by very much more than the 15 % caused by constant emission operation. In other words, valve life is being wasted if there is excess emission at any time during life.

Mr. Grinstead also raises the question of differences in secondary emission between molybdenum and tungsten grids. It is our general experience that, in the secondary emission region, i.e. grid voltages below about 400 volts, greater secondary emission is obtained from molybdenum grids. This agrees with results of measurements of the secondary emission coefficient of molybdenum and tungsten.*

No relationship between X-ray photoelectric emission and secondary electron emission is known.

In reply to Mr. McLarty, we are quite confident that air-cooling, even of valves of the largest size, will be practicable.

With regard to the magnetic modulation of characteristics by alternating filament current, a distinction must be made between the effects as they exist in the valve, and the resultant " hum " which actually appears in the output circuit. For any given type of valve the hum level varies very widely according to the circuit conditions. Thus for the C.A.M.5 type, in which the modulation has been measured oscillographically, we understand that in the case quoted by Mr. McLarty the circuit conditions did not require the valve to amplify as well as to modulate.

In reply to Mr. Turner on the question of straight-line characteristics, the several controlling factors are not numerically investigated in every case. Nevertheless

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* *L'Onde Électrique*, 1937 vol. 16, p. 516.

Fig. 10 is, as stated, representative of the great majority of types, including triodes, so that the three principles which he enumerates are now suitably included in the general tradition of valve design. Straightness of a large proportion of the characteristic is not, in general, necessary, though it is usually desirable that the upper part should be straight; the form of the lower part is also important. Thus, for instance, either for a single amplifier valve with a tuned load circuit of moderately high "Q," or for valves in push-pull with any type of load, a linear relation between input voltage and output current can be obtained from an individual dynamical characteristic which is parabolic in the lower part and passes upwards into a tangent line; in fact the characteristics of most valves approximate to this form. More generally, however, it can be shown that a linear response can be obtained from any characteristic which satisfies the condition that, the output current being expressed as a series of powers of the input voltage, this series shall contain no odd powers except the first. A parabola satisfies this condition, but since the apex is at the cut-off point, and the curve cannot rise again on the negative side of cut-off, the parabola has to pass upwards into a tangent line if the operating range is not to be unduly restricted. A hyperbola with an asymptote coinciding with the voltage axis also satisfies the condition. This shows that linearity is possible even with an extended "tail" at the foot of the characteristic.

With reference to Fig. 11 and its theoretical basis, what we have tried to do is to show that there are adequate theoretical reasons for supposing that all current values are, under the conditions outlined, multiplied by a common factor n, and all voltage values by another common factor $n^{\frac{2}{3}}$.

That the result of these multiplications is a diagonal displacement of a curve in Fig. 11 is surely a basic geometrical property of the log/log method of plotting. The only hiatus in the argument that we have been able to find is that we have omitted to state that the constant of the $\frac{3}{2}$ power relation between voltage and space-charge-controlled current for each element of the surface is independent of the temperature.

We hope that the matter will now be clear, for the principle is a very useful one; we may add, for instance, that the power-handling capacity of the valve is multiplied by $n^{\frac{5}{3}}$.

As to the heating current "leaving the positive end" of the filament, the electrons which constitute the current certainly do so, and we remain incorrigible in discarding convention and thinking of current in that way. If one does so it is at least always clear that the negative end must be the hotter.

We agree that we should have included a scale of dimensions in Fig. 6; actually, the external diameter of the anode is 8 inches.

The evaporated tungsten mainly comes down as a deposit on the anode, usually smooth but sometimes flaky, but in earlier days some of it used to appear as hairy or plate-shaped crystalline growths on the filament itself. In reply to Mr. Grinstead, for the check observations included in Fig. 7(*b*) it was not necessary to use the intermittent method, because with alternating-current filament heating a space-current of some 12 % of the heating current is not sufficient to affect the distribution of filament temperature.

In reply to Mr. Moullin about the Schottky effect, our difficulty in assessing the value of Mr. Gill's work is briefly as follows: Schottky's treatment starts from a very simple basis, the counteraction of the image force by the external field, and arrives at a relation involving a function of a certain exponential form including certain natural constants. Now exponential functions do not of their nature lend themselves very well to exact numerical verification—we all know what log plots are—but in this case the values of the constants have been verified with great exactitude, e.g. by de Bruyne in the work cited. This is our real stumbling-block, for Mr. Gill proposes to substitute an empirical function supported only by a very tentative theory which gives no clue to the values to be expected for his constants. We feel that cases such as oxide-coated or thoriated surfaces, in which Schottky's theory fails, and may be expected to fail since they violate his initial assumption of a smooth homogeneous surface, suggest that the Schottky theory should be suitably extended rather than discarded in favour of an alternative with an entirely different basis.

Mr. Turner asks for a definition of "equilibrium pressure." This descriptive term is used because the pressure as measured represents a balance between the liberation of gas from the electrodes, etc., and the disappearance of gas by clean-up.

The fixation of gas referred to by Messrs. Ponte and Warnecke is what we have called "clean-up." The effect is particularly noticeable in a valve having a large clean-up capability and a little free gas. When a high anode voltage is applied and a little anode current is allowed to flow, the free gas rapidly disappears and the backlash ratio approaches the photoelectric limit. This treatment was applied to the valve before the curve of Fig. 2 was taken.

During our work on the X-ray photoelectric effect no investigation was made into the *K* and *L* absorption limits of molybdenum, or into the variation of intensity of the soft X-rays with exciting voltage. Further information on these points can, however, be obtained from the original work of Richardson and Robertson.*

Regarding the magnitude of the photoelectric component of grid current: this, even for the largest valve, is but a few hundred microamperes and is small compared with the grid current which flows during normal operating conditions. Thermionic grid emission may be many orders of magnitude greater than the photoelectric component, but it was checked that Curves A and B of Fig. 5 were in no way influenced by thermionic emission from the grid. For the valves described in the paper the operating grid temperature is below that at which thermionic emission from the metal itself occurs, and it is only in extremely rare instances, such as that mentioned by Mr. Walker, when a contaminant of the sort described by Messrs. Ponte and Warnecke finds its way to the grid surface that thermionic emission from the grid is encountered.

The disadvantage of de-gassing anodes by bombardment is that clean-up inevitably proceeds simultaneously and some of the gas evolved by the anode is cleaned-up in other parts of the valve. In some cases this may not be a serious disadvantage, but it is our experience that perfectly satisfactory valves can be produced without anode bombardment.

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* *Proceedings of the Royal Society*, A, 1927, vol. 115, p. 280, and 1929, vol. 124, p. 188.

Before considering the further points raised by Messrs. Ponte and Warnecke we should like to mention their recent publication,[*] which is on similar lines to the present paper.

Regarding the relative merits of different forms of cathode seals, it is, to some extent, a matter of personal experience and technique whether the water-cooled rod type of seal or the cone type described in the paper is preferred. We do not agree, however, that bad contacts can be cited as a defect of the latter type of seal. In our experience this fault is entirely negligible.

No comparison with chrome-iron seals was intended in the paper. We have never used this type of seal and therefore have no experience of it.

Reference has been made to the circular grid seal as used for type C.A.T.17. This seal, of course, forms a very convenient mounting for the grid system, and as far as distribution of high-frequency current is concerned it is at least as good as two side seals of conventional type. A great advantage of the circular seal described is the large surface area available for contact. This is important when currents of the order of 100 amperes are being handled under normal operation, and, owing to the large dimensions of the seal, it is capable of dealing with current overloads due to mistuning, etc. The temperature-rise of these seals in operation is small, and the temperatures reached during normal evacuation are sufficient to ensure that no further gas is evolved under running conditions. We agree with Messrs. Ponte and Warnecke that it is often desirable to employ special methods of de-gassing the grid seals and leads of short-wave valves, particularly when the assembly is not so lavishly designed as that of type C.A.T.17.

In reply to Mr. Greig, constructions in which the high-frequency current is by-passed from the grid supports by flexible low-inductance strips have been considered, and in fact have recently been adopted experimentally. The effect of the inductance of the cathode leads in conventional valves usually becomes apparent before transit-time effects. Fortunately, however, the cathode may be brought to earth potential by means external to the valve, e.g. by the use of a half-wavelength line between cathode and earth. On this account it has not so far been found necessary to reduce the inductance of the internal filament leads.

We do not understand the remarks of Messrs. Ponte and Warnecke on the subject of low-μ grids. We have called a μ value of less than 10 "low" and such grids are usually made in the form of parallel rods. Valves of this sort, with μ as low as 6·5, are quite commonly in operation under Class B high-frequency conditions where the grid voltage becomes positive during part of the cycle.

Where we use grooved anodes to increase the effective area the jacket is brought in so close that there is no possibility of the water-flow avoiding the grooves, and we do not find that deposits form more rapidly in the grooves than on what remains of the outer surface of the anode.

We do not agree that channelling of the inside of the anode is better than blackening. On the contrary, although we use channelling in some cases its effect is less than we had hoped, and blackening has a much greater effect. Choice of a suitable material avoids the other difficulties mentioned.

Turning now to the division of current between anode and grid, we should first make clear, as we may not have done in the paper, that the observations on which diagrams such as Fig. 13 are based always include readings taken with anode and grid at the same potential. For most purposes, therefore, the "straight line" is used for interpolations within a not-very-large gap. Our principal concern has been with I_g as a correction to extrapolated values of I_a, rather than with estimation of the I_g characteristic itself or with the grid dissipation. We have thus not given the question so much study as Messrs. Ponte and Warnecke.[*]

However, we should like to put forward some comments in reply to their queries. In Lange's work the factor $\sqrt{(E_g/E_a)}$ appears at an early stage after comparison of the respective fields and space-charge distributions along lines between cathode and grid member and between cathode and grid space. The actual arguments by means of which he introduces it appear to us to be fallacious, though we would not deny that there may be some relation between the voltage-ratio and the space-charge. It has, however, been pointed out long ago by J. J. Thomson,[†] and more recently confirmed by Treloar,[‡] that this same factor $\sqrt{(E_g/E_a)}$ must also appear when the deflection of the electrons by a charged grid wire is considered. This aspect is dealt with by Lange by means of his "correction angle" β, but is not made the subject of analytical treatment. It seems, therefore, that his simple formula, as quoted by Messrs. Ponte and Warnecke, needs to be extended so as to include both of the effects of the factor $\sqrt{(E_g/E_a)}$. The footnote to M. Warnecke's paper seems to imply that they are of the same opinion. Further, it seems that Lange's treatment of the space-charge of primary electrons in the neighbourhood of the grid wires, which is what we named the "incoming space-charge," only includes the effects of those electrons which actually strike the grid. Other electrons, however, which have been retarded by the local grid field but are deflected so as not quite to strike the wire, must also contribute to the local space-charge and increase its effect. It may be remarked here that in large valves the total surface area of the wires in the active parts of the grid is only about three times the cathode area, and the deflection of electrons by a negatively charged grid must effectively reduce to a small fraction of the total the main area bombarded. The current density in the bombarded areas is therefore quite high, and the combined effect of the bombarding electrons and their neighbours may therefore be considerable. It was for this reason that we were inclined to ascribe the suppression of secondary emission by local reversal of the field to this "incoming space-charge." We confess, however, that we have omitted to consider the outward movement of the main cathode space-charge, as discussed by Lange, as a contributory factor. Finally, in all cases where it is known that the grid current is influenced by the magnetic field of the filament current to the extent that it is in large valves no purely electrostatic theory, such as the suggested extension of Lange's, can be regarded as providing a sound basis without careful consideration. This subject is difficult and complicated and we are much indebted to Messrs. Ponte and Warnecke for their observations.

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[*] *Bulletin de la Société Française Radio-électrique*, 1938, vol. 12, p. 29.

[*] *Loc. cit.* (concluding footnote).

[†] *Journal I.E.E.*, 1920, vol. 58, p. 682.

[‡] *Proceedings of the Physical Society*, 1936, vol. 48, p. 48.

We are indebted to Mr. Gibson and Mr. Pennington for their interesting contributions to the discussion. We find they have left us nothing to say in reply.

ELECTRICAL INTERFERENCE WITH RADIO RECEPTION*

By A. J. GILL, B.Sc.(Eng.), Member, and S. WHITEHEAD, M.A., Ph.D., Associate Member.

(*Paper first received* 21*st October*, 1937, *and in final form* 29*th July*, 1938; *read before the* Wireless Section, 6*th April*, 1938.)

SUMMARY

The paper describes the method of assessment of the interference to radio reception from electrical equipment, and determines the level to which such interference must be reduced to permit satisfactory service.

The methods of achieving this result are described for the various classes of interfering equipment. Although mainly directed towards the protection of broadcast reception, the principles described apply equally to other radio communication services.

(1) GENERAL

(a) Nature of Interference: Signal/Noise Ratio

Any change in the electrical conditions of a circuit gives rise to a spectrum of components of current or voltage, and the more abrupt the change the higher will be the frequency to which these components extend.[†] Nearly all classes of electrical equipment are subject to these rapid changes, while, in addition, the apparently continuous wave-shapes of current and voltage possess series of harmonics which may extend into the radio-frequency region. Thus nearly every item of electrical equipment and apparatus may, from the present aspect, be regarded as a potential source of radio-frequency energy. This energy may be transmitted to the aerial of a radio receiver, producing a radio-frequency interfering voltage v at its input terminals with corresponding unpleasant sounds in the loud-speaker, or other reproducing apparatus.

The interfering voltage v is received simultaneously with a modulated carrier which, if conditions are to be tolerable, must be much greater than v, the effect of the latter being the more pronounced in the passages of lighter modulation. Thus the conditions of reception to be assumed in evaluating v are those corresponding roughly to the presence of a large unmodulated carrier. A suitable output voltmeter must be such that, if connected in parallel with the receiver output in the conditions defined, it gives equal indications when interfering voltages of different types, applied to the receiver, produce noises in the loud-speaker judged by listeners as of equal disturbing effect on a programme which may be superimposed on the noise. Such a meter, as regards pure continuous notes, would follow more or less the response/frequency characteristic of the ear. A frequency-weighting curve of this type is shown in Fig. 1.[‡] This subject has received extensive study by the International Advisory Committee for Long-distance Telephone Communication in connection with line communication, but in telephone practice the properties of the receiver mainly determine the phenomena. In radio communication the noises are usually either discontinuous or of such complex character that the charging and discharging times are of more consequence than the frequency response.[†] As a result of the work of the Comité International Spécial des Perturbations Radiophoniques[‡] (C.I.S.P.R.) a voltmeter fulfilling the above conditions has been specified as a peak voltmeter of linear scale over the working range having a charging time of 1 millisec., a discharging time of 160 millisec., and an indicating-needle time-constant of 160 millisec. The instrument is critically damped. A frequency filter added to give a response similar to that in Fig. 1 has been shown to add little to the accuracy.

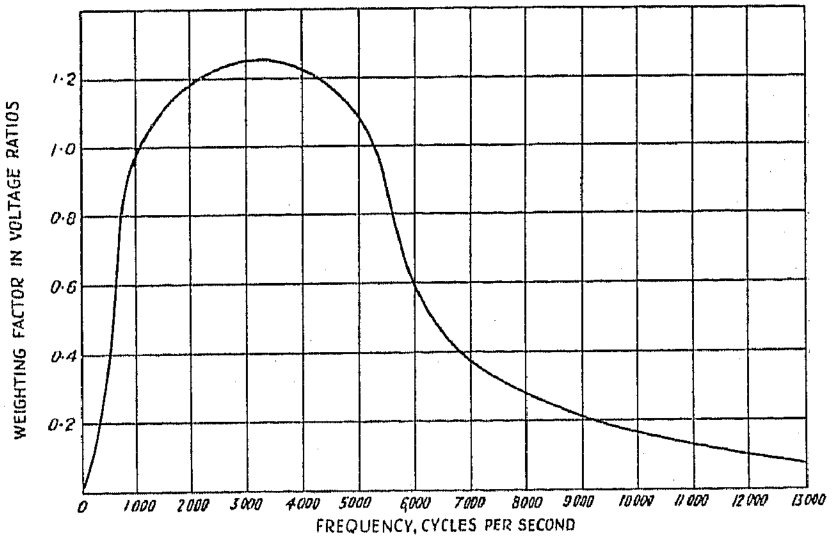

Fig. 1.—Weighting curve for broadcast psophometer (C.C.I.F., Budapest, 1934).

This specification does not deal with the effect of duration and frequency of repetition of the interference, features which are often very difficult to define for a given type of disturbance. Accordingly it is usually satisfactory to take the largest values observed which occur at all frequently. The standard adopted by the B.S.I. is to take any steady reading if it is maintained beyond 10 sec. or exceeds on the average a duration of 1 sec. per hour, or any interference, however short, if repeated oftener than once in 10 minutes. A large tolerance must in any case be allowed in respect of accuracy of measurement. Regularly-repeated interference will give indications corresponding to a single impulse if the interval between successive impulses is greater than about 0·2 sec.

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[*] Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 345.

[†] For example, a square-topped pulse of time-of-passage t gives rise to a spectrum of an intensity, in the band comprised between the wavelengths λ and $(\lambda + d\lambda)$, which is jointly proportional to $\sin^2(2\pi ct/\lambda)$ and to the band width $d\lambda$, where c is the velocity of light. Any given pulse can be derived by summation of such elements.

[‡] See Proceedings of C.C.I.F., 1934, vol. 4, p. 258 (Budapest Meeting).

[†] K. Muller and U. Steudel: *Veröffentlichungen aus dem Gebiet der Nachrichtentechnik*, No. 2, 1935.

[‡] See Reports (RI)3 and (RI)4 of the I.E.C.; also B.S.S. No. 727—1937.

On the other hand, if the interval is less than about 0·1 sec. the voltmeter will give a reading approaching closer to the peak value of the interference. If the individual impulses are of a rapidly transient character the increase in indication of the voltmeter, as the impulses are repeated with a greater frequency, may be much greater than the corresponding increase of subjective impression. This source of error is rarely important with interference on the normal broadcast wavelengths, but with ignition systems the range of spark frequency may include the frequencies corresponding to the discharge time-constant of the voltmeter. The difficulty is conveniently overcome by referring the interference to a "normal" spark frequency.

The greater the proportion of soft or lightly-modulated passages in a programme the greater will be the disturbing effect on it of a given interference. Speech and music differ also in intrinsic liability to disturbance. A number of tests were made by the C.I.S.P.R. in which various observers adjusted the level of different types of interference, superposed on a range of representative programmes, until it was judged just tolerable. The interference was measured by the voltmeter already specified, and the programmes by a peak-voltage indicator. The signal(programme)-to-noise ratio so determined as corresponding to the limit of toleration was then found to be substantially independent of the observer, type of programme, and type of noise, provided that the maximum level, i.e. the peak of the dynamic characteristic* of the programme, was taken. In these circumstances the limiting signal/noise ratio was found to be 40 db., the same ratio as that applying to interference between neighbouring transmissions. Now the peak of the dynamic characteristic corresponds to the peak modulation of the transmission In general broadcasting practice the peak modulation level is usually taken as 80 %, so that the general criterion of "freedom from interference" is that the noise due to the interference, measured by the voltmeter specified, should be at least 40 db. below the level arising from the transmission to be protected when having a degree of modulation of 80 %. The same principles will apply to the sound transmission accompanying television. The requirements for the vision transmission have not yet been decided. Experience (mentioned later) so far suggests that the sound is more sensitive to interference than the vision; so that, from the point of view of interference, television may be conservatively regarded for the present as an ultra-short-wave broadcast.

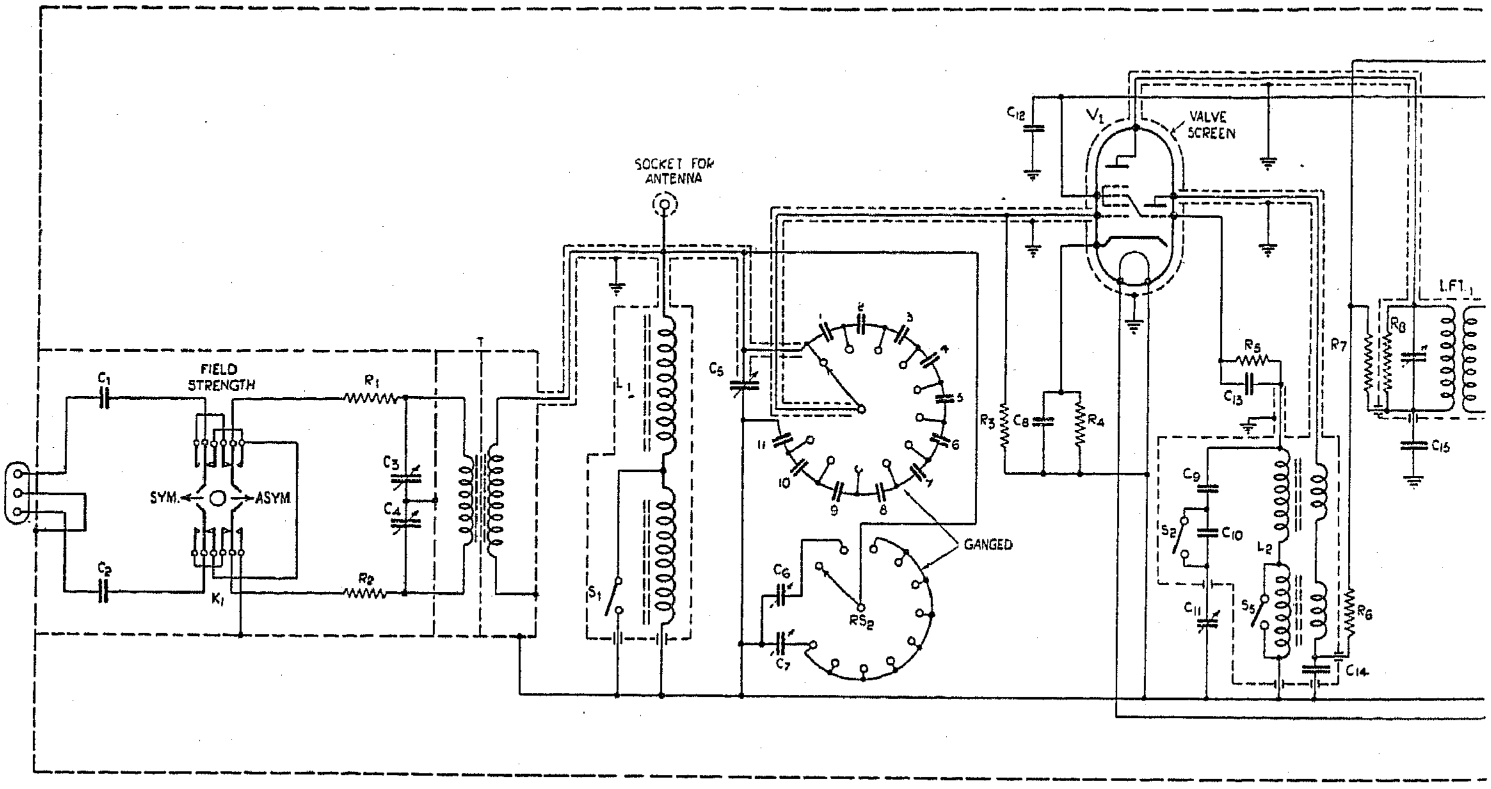

Fig. 2.—G.P.O. portable radio-interference measuring set.

(b) Measurement of Interference

Let it be assumed that an ideal receiver selects a band of frequencies, rectifies the resultant in the manner known as linear, and rejects all except a certain band of low frequencies, but otherwise has no action upon the input signal apart from a constant amplification of the frequencies selected. Then it may be shown (as in the Appendix) that if the incoming signal is

$$P(1 + M \cos pt) \cos \omega t + \Sigma v_n \sin (\omega_n t + \phi_n) \quad . \quad (1)$$

then the low-frequency output is, in certain conditions,

$$V = \text{Const.} \times \left\{ MP \cos pt + \Sigma v_n \sin \left[(\omega_n - \omega) t + \phi_n \right] \right\} \quad (2)$$

We may regard P as the intensity of the carrier, and $M \cos pt$ as its degree and type of modulation, while $v_n \sin (\omega_n t + \phi_n)$ is a radio-frequency interfering voltage lying in the band accepted by the receiver.* Thus

$$\text{Signal/noise ratio} = MP/\Sigma v_n \quad . \quad . \quad . \quad (3)$$

both quantities being measured by a meter of suitable response. From the preceding section, M must be taken as 0·8 (i.e. 80 % modulation) in determining whether the interference is tolerable. Thus the signal/noise ratio is numerically equivalent to 0·8 times the ratio of the carrier amplitude to the integral of the high-frequency interference spectrum over the band width selected, provided this integration is made in such a way as to take account of the time-constants of the voltmeter specified. This leads to the important conclusion that the relative interfering effect can be expressed as a property of its high-frequency spectrum (selected and integrated in a defined manner) independently of the signal which is disturbed. This interfering quality should be measurable, in the absence of a signal, by a suitable high-frequency (h.f.) voltmeter. If the noise voltmeter previously defined had been a simple peak voltmeter a similar h.f. peak voltmeter would have been an equivalent noise-indicator. Nevertheless, it is shown in the Appendix that an h.f. voltmeter of the same time-constants should give indications approximately equivalent to those given by the noise voltmeter defined.

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* Subject to the maximum level being attained or exceeded for reasonable periods, e.g. at least about 5 % of the programme time.

* Assuming that the h.f. band-width corresponds to, or is narrower than the l.f. band-width.

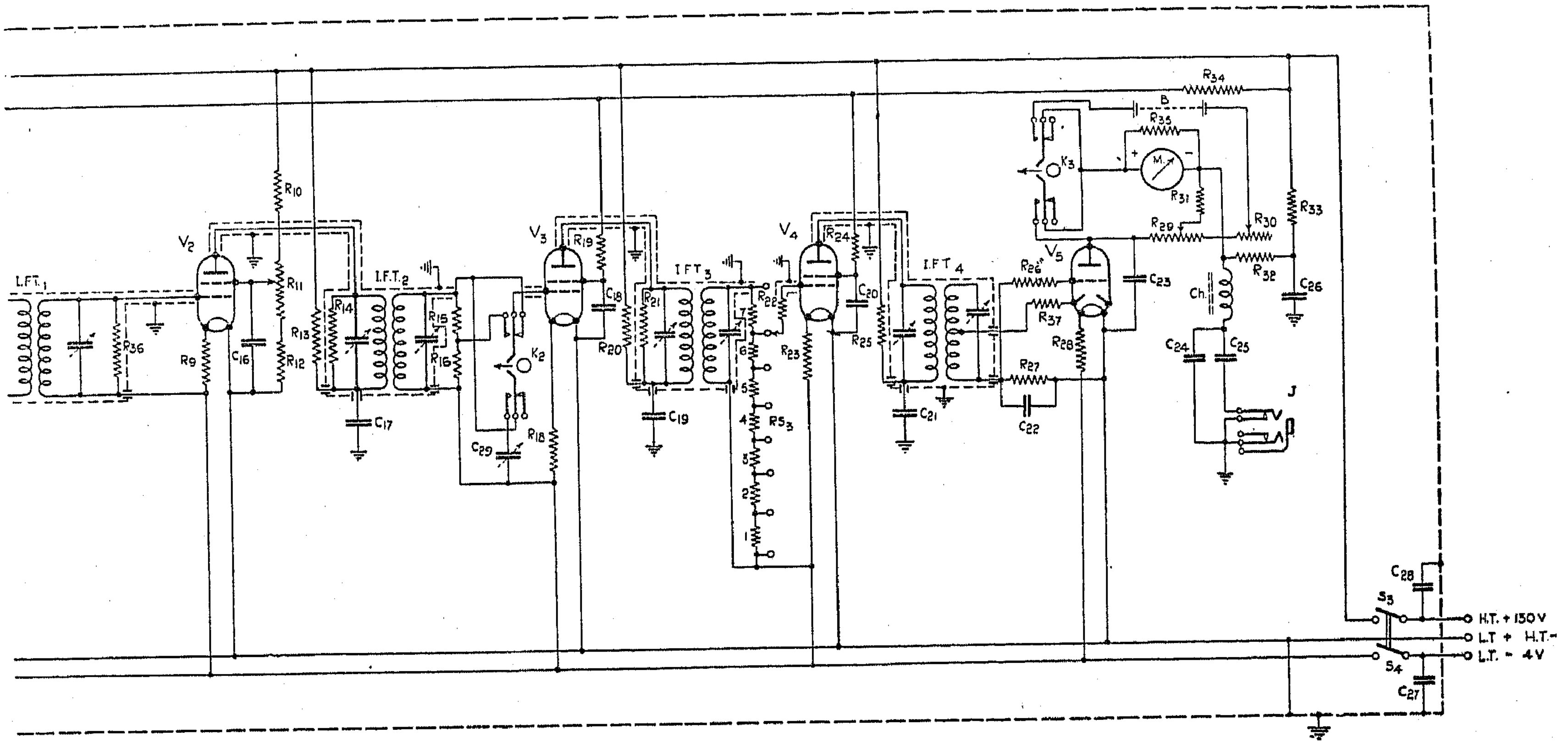

Fig. 2 (continued).

The validity of the foregoing analysis requires the following conditions in the ideal receiver to which it refers:—

(i) The receiver is of perfect linearity and fidelity.

(ii) The band width is selected by the h.f. stages, the low-frequency (l.f.) stages imposing only an equivalent limitation such that terms of frequency greater than the maximum modulation frequency permitted by the band width are eliminated.

(iii) The signal carrier is very large compared with the interference.

(iv) The reaction of the sidebands of the signal on the interference is neglected.

It is shown in the Appendix that in the absence of the signal the low-frequency output of the receiver is given approximately by:—

$$V = \text{Const.} \times \Sigma v_n \sin\left[(\omega_n - \omega')t + \phi_n\right] \quad . \quad (4)$$

where ω' is the frequency of rectification, both ω' and ϕ_n being actually variable. Although (4) obviously differs considerably from the second terms of (2) if n is small, it is shown in the Appendix that when n is large (i.e. when the interference spectrum tends to be continuous, as is usual in practice) the maximum value of (4) is probably equal to the maximum value of the interference terms in (2), so that the carrier may often be omitted in measuring the interference, without great error.

The average listener's receiver falls short of the ideal, but two possible defects are easy to assess. The standard band-width adopted is that corresponding to the transmission of frequency-displacements from 100 to 4 500 cycles per sec. within the following tolerances: + 1·5 db. and − 20 db. with respect to the response of the carrier frequency for frequencies displaced less than 100 cycles per sec.; + 1·5 db. and − 6 db. from 100 to 200 cycles; + 1·5 db. and − 1·5 db. from 200 to 2 000 cycles; + 1·5 db. and − 6 db. from 2 000 to 4 000 cycles; + 1·5 db. and − 20 db. from 4 000 to 5 000 cycles; − 6 db. and − 20 db. from 5 000 to 10 000 cycles; response less than − 20 db. for frequencies displaced more than 10 000 cycles. When completed, the ideal characteristic will presumably be that of the standard apparatus of the C.I.S.P.R. now in construction. If the actual band-width differs from this, then the interference, if of the continuous-spectrum type, will vary proportionately with the integral taken over the selectivity (response/frequency) curve, while the broadcast signal will usually be affected to a smaller extent.[*] With a narrow band-width the effects tend to be the same for both, and the signal/noise ratio is not affected. If the rectification is non-linear the signal/noise ratio is still approximately given by (3). Mumford[†] has shown that,

[*] A narrow band-width diminishes the broadcast sidebands, while a wide band-width introduces demodulation effects.

[†] A. H. Mumford and J. L. Howard: Radio Report No. 277, 1934 (P. O. Engineering Department). See also A. H. Mumford and H. Stanesby: Radio Report No. 300, 1934 (P.O.E.D.); and J. W. Alexander: *Hochfrequenztechnik und Elektroakustik*, 1932, vol. 40, p. 82.

neglecting the signal sidebands and intermodulation products of the interference, this conclusion is strictly true as regards r.m.s. values in square-law rectification, and his analysis will lead also to the more general indication that the conclusion is roughly true as regards the modified peak values, provided the rectification follows a power law with index between 1 and 2.

It is possible by the means described to make an absolute measurement of the interfering quality of the voltage appearing at the listener's aerial, but to characterize the disturbing effect of a source of interference it is necessary to consider the coupling between the source and the listener. When the interference is directly radiated, its field strength may be found, using a calibrated aerial in conjunction with a measuring set of the defined characteristics, and may be specified at the minimum distance at which a listener's aerial may reasonably be situated; or the field-strength distribution may be plotted and directly compared with the assumed or measured broadcast field-strength. As regards indoor aerials, the broadcast field and, where possible, the interfering field, in free space in the immediate neighbourhood of the listener are considered. Often, however, the interference is partly conducted, as by the electric mains, and then radiated therefrom to the listener's aerial. It is then sometimes possible to determine a statistical average field-strength of the radiating system when excited by the various sources normally operating.[*] The more important instance of the domestic electric mains[†] does not permit the satisfactory definition of a field strength and accordingly the ratio A of the interfering voltage at the listener's aerial to the interfering voltage at the terminals of the disturbing item is measured and known as the "coupling factor." The effective height h of the listener's aerial referred to the broadcast field in the neighbouring free space is also measured. Then if V is the interfering voltage of the item and ϵ the broadcast field the condition for good reception is that—

$$V \leq 0 \cdot 008 \frac{h\epsilon}{A} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

A statistical analysis of the quantity (h/A) permits the determination of the limiting value of V, and the interfering quality of the item can be characterized by the measured value of V. It is necessary to consider both the voltage which may exist between the terminals of the item and the voltages between the terminals, which may be connected together for this measurement, and earth.[‡] The former voltage is known as the "symmetrical voltage" and the latter as the "asymmetrical voltage." In addition, the disturbing voltage is profoundly influenced by the external h.f. impedance applied to the item through the system connected thereto. The conditions, in these respects, under which the item is tested must therefore be carefully defined for each type of disturbing source.

(c) Comparison of Various Methods

The foregoing principles lead to three types of measuring equipment. In the first method an artificial carrier may be injected at the input or at the detector, and the special voltmeter defined measures the l.f. output. This method is employed in the Siemens and Halske apparatus,[*] which comprises a 2-stage h.f. amplifier tuned to the injected artificial carrier, a square-law detector, which functions linearly owing to the large carrier, and an l.f. amplifier and special voltmeter. A simplified version employs an oscillating detector and a voltmeter consisting of a single valve. A feature of the standard apparatus which seems open to question is that the band width is partly selected in the l.f. stages after detection.

In the second method the special voltmeter is used in conjunction with a suitably selective h.f. amplifier as a high-frequency valve-voltmeter. This method has been adopted by the G.P.O. and the E.R.A., although a number of tests have been made by the previous method. Besides the use of standard field-strength-measuring equipment, the G.P.O. have designed a portable testing set, shown in Fig. 2 and described in B.S.S. No. 727—1937. It comprises a superheterodyne amplifier and incorporates means for calibration, *in situ*, by observing the indication due to the thermal agitation in the first tuned circuit. A similar set having a range from 12 to 2 000 m. has recently been constructed. The E.R.A. set for the broadcast ranges (Fig. 3) employs a buffer-amplifier type of tuned amplifier. Means are incorporated for checking battery voltages and currents, and the apparatus has been found sufficiently constant in use to need only an occasional recalibration by a standard signal. Both the G.P.O. set and the E.R.A. set employ a vertical-rod aerial in measuring the field strengths and are adapted for measuring interfering voltages by means of a balanced and screened input transformer. The E.R.A. short-wave measuring set (Fig. 4) is only adapted for measuring field-strengths using a frame or vertical-dipole aerial. It is similar to the G.P.O. apparatus in employing the superheterodyne principle (the intermediate frequency is 1 Mc.) but is directly calibrated for field strength by the radiation from a given loop carrying a known current at a standard distance. Recently the h.f. method has been adopted in the standard C.I.S.P.R. apparatus now in construction, and has also been used in the U.S.A.[†]

The third method, the measurement of the l.f. output of the interference in absence of a carrier, is employed in less accurate, in comparative, and in qualitative, tests, and can be applied to a calibrated radio receiver equipped with a suitable output voltmeter. The N.E.L.A.[‡] in America standardized this method, specifying the selectivity and fidelity curves. The h.f. band-width so defined is rather less than 9 kc., but the l.f. fidelity corresponds roughly to the C.I.S.P.R. definition. No acoustic frequency-weighting is employed, and the integration time, determined by the meter needle, is of the order of 0·3 sec. A transfer standard method is adopted, a multivibrator (120 cycles per sec. fundamental frequency) being adjusted to give the same reading as the interference. The output of the multivibrator is cali-

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[*] The method followed with traction and communication systems, etc.

[†] F. Eppen and K. Müller: *Elektrische Nachrichten-technik*, 1934, vol. 2, p. 257.

[‡] K. Müller: *Veröffentlichungen aus dem Gebiet Machr. Techn.*, 1934, vol. 4, p. 139; W. Wild: *Elektrotechnische Zeitschrift*, 1933, vol. 54, pp. 149, 172.

[*] Siemens and Halske Publication No. 491 C.; W. Wild: *loc. cit.*; J. Schmied: *Elektrotechnik und Maschinenbau*, 1935, vol. 53, p. 373.

[†] F. O. McMillan and H. G. Barnett: *Transactions of the American I.E.E.*, 1935, vol. 54, p. 857.

[‡] National Electric Light Association Reports Nos. 32 and 33; *General Electric Review*, 1933, vol. 36, p. 201.

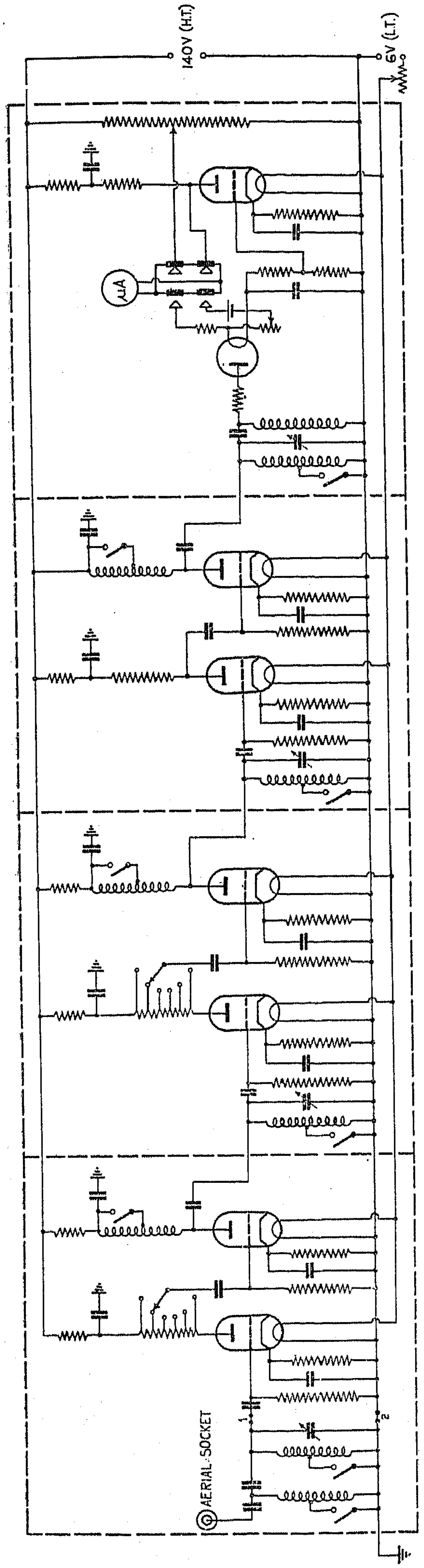

Fig. 3.—E.R.A. broadcast-frequency interference-measuring set.

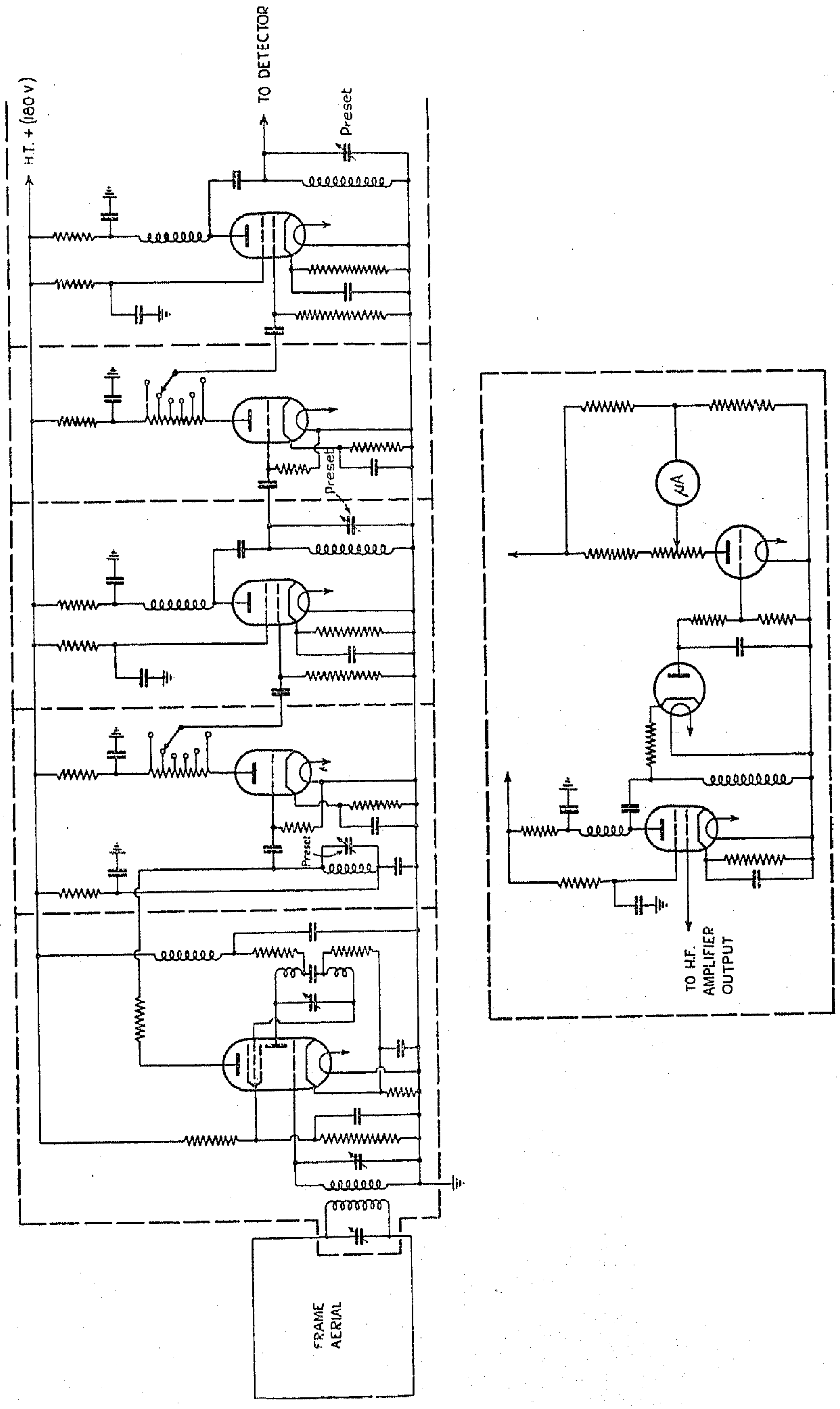

Fig. 4.—E.R.A. short-wave measuring set.

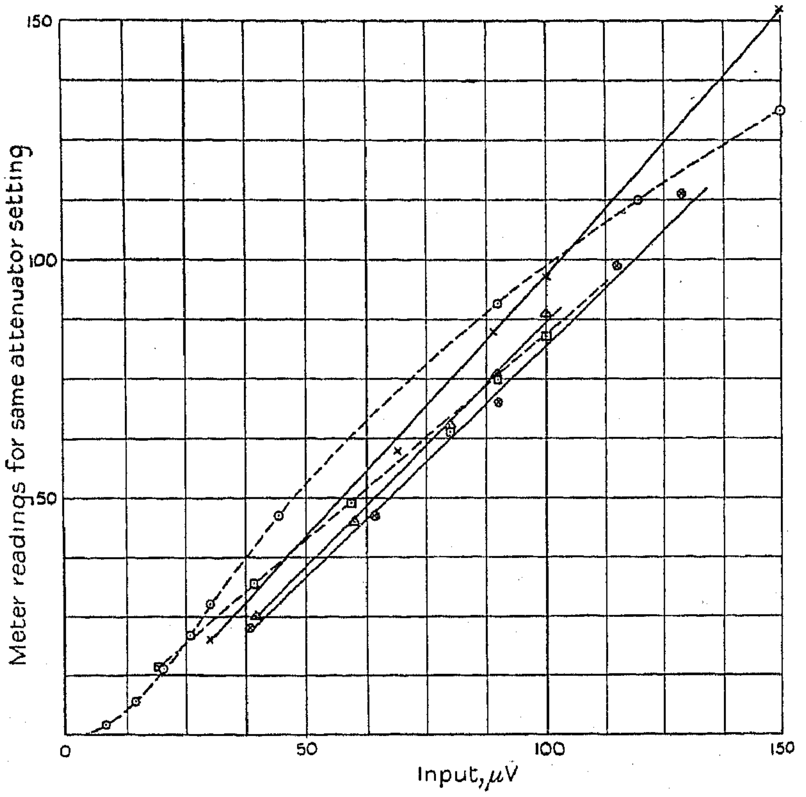

Fig. 5.—Effect of simple signals with different methods of measurement.

	First method	Second method	Third method
Signal	L.F. voltmeter with carrier	H.F. voltmeter	L.F. voltmeter without carrier
Single frequency	– – ⊡ – – 0·4 kc. from carrier	——X——In tune ——△—— Displaced 0·4 kc. from peak	No reading
Wave modulated 30 % at 0·4 kc.	Same as third method in this case	—⊗—Corrected for crest value	– – ⊙ – – Corrected for modulation percentage

brated in terms of a standard signal modulated at 400 cycles per sec. to a depth of 40 %. Similar equipment, with the addition of an acoustic filter adjusted to Fig. 1, was used in France,* but the C.I.S.P.R. methods are now followed.

The E.R.A. receiver was arranged to make measurements according to any of the three methods. Fig. 5 shows that the reactions to standard signals follow the elementary theory. An unmodulated wave gives the same readings with the first two methods and no reading with the third. If a wave modulated to a depth M is applied the first method gives $(1 + M)$ times the carrier (peak h.f. voltage), while the second and third methods are identical and give the modulation M. The readings are corrected by the factors mentioned, and the agreement is within the differences caused by the selectivity, curvature, and lack of perfect linearity in rectification. Similar comparisons between the first and second methods were made by the G.P.O. and E.R.A. with different sources of interference, as shown in Table 1.

Thus the first and second methods, which are theoretically equivalent, are also equivalent in practice, allowing for the difficulty of reproducing accurately the interference tested. The complexity of the interference employed in these tests is evidenced by the fact that an average h.f. voltmeter gave results differing from those referred to in Table 1 to a mean extent of 8 db., and in some cases by 10–20 db. Comparison (49 tests) between the G.P.O. and E.R.A. sets, both following the second method, gave a mean difference of 0·12 db. and a maximum difference of 4 db. A similar comparison was made between the G.P.O. and the Siemens and Halske sets, the latter embodying the first method. The mean difference was 3 db. and the maximum 6 db.

Provided the characteristic of the measuring apparatus is accurately linear and the interference has a continuous and fairly uniform spectrum, the presence of the carrier, wherein the first method differs from the third, has not a very great effect. This was verified both with artificial and with normally-radiated carriers in tests on interference from trolley-buses and also (artificial carrier only) in tests on domestic items. In any case, the ratio of unsuppressed to suppressed levels is less affected by the carrier than absolute values, so that earlier work on such ratios by the third method remains substantially valid.

Table 1

Tests made by	Source of interference	Difference* (db.)	
		Max.	Mean
E.R.A.	(1) Buzzer of telephone noise unit	+ 2	+ 1
	(2) Fractional-h.p. motor	+ 4	+ 3
	(3) Magneto ignition system	− 5	− 1
	(4) Dynamo and commutator (square impulses)	− 2	− 0·5
	(5) N.E.L.A. multivibrator	+ 4	0
G.P.O.	(6) Vacuum-cleaner motors (fractional-h.p. universal type) . .	+ 2	+ 1·5
	(7) Wheatstone transmitter run at 10 speeds varying between 1 and 90 pulses per sec.	+ 3	+ 1

* A plus sign means that the l.f. voltmeter with carrier gave higher readings than the h.f. voltmeter without carrier.

The importance of polarization in interference measurements is mainly confined to frequencies above 20 Mc., where also regard must be paid to reflection, wave-front distortion, etc. A reasonable degree of consistency has been obtained with the E.R.A. receiver by orienting the frame to give the maximum reading and by taking a mean for several points in the neighbourhood of the test position. The signal/noise ratio has been found in tests on various types of interference to be the same for a vertical-dipole as for a frame aerial. Table 2 gives the results of some subjective tests carried out on the B.B.C. television service.

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* H. SUBRA: *Annales des Postes, Télégraphes, et Téléphones*, 1935, vol. 24, p. 368.

Table 2

Type of interference	Tolerable signal/noise ratio, db.		Remarks
	Sound	Vision	
Valve electrotherapy apparatus (unsmoothed a.c.)	40	16–17	The h.f. band-widths of the receivers employed were 60 kc. for the sound transmission and 4 Mc. for the vision transmission. Ratio of interfering field by frame aerial to value by dipole aerial was + 0·6 db. at 41·5 Mc. and − 2·5 db. at 45 Mc.
Ignition systems	30–43	21–24	
Contactors	34	33	

(2) SOURCES OF INTERFERENCE

(a) Radiated Interference

(i) Trolley-buses.

Although included for convenience in the present Section, the interference arising from the operation of a trolley-bus is mainly radiated from the overhead contact wires along which the disturbances are propagated. The direct radiation from the vehicle itself is rapidly attenuated with distance and is usually less than 10 μV per metre beyond about 20 yards. The radiation from the overhead system extends along the wires for a considerable distance, sometimes along the whole system, and may exceed 10 μV per metre at distances from the track up to ½ or 1 mile. Fig. 6 shows the variation, with distance from the track, of the field strength of the interference caused by a vehicle passing opposite the test point. In the tests shown in Fig. 7 the interference did not vary greatly with frequency, but differences in this respect are encountered in different systems. The sources of interference on a trolley-bus are the collectors, the main contactors supplying and controlling the motors, the relay circuits controlled by the driver which operate the contactors, the driving motor, the brake (exhauster or compressor) motor, and the motor-generator (if present) for lighting. It is desirable also to distinguish between rheostatic braking, where the motors during deceleration are disconnected from the supply and dissipate the brake energy in a rheostat, and regenerative braking, where energy is fed back into the line. Some recent types of trolley-bus combine the two forms, rheostatic braking being employed at low speeds. The interference from trolley-buses is assessed on a field-strength basis and is usually referred either to the maximum recurrent disturbances observed at a distance of 10 yd. from the track, or to observations made underneath the contact wires, at a distance of 40 ft. from the vehicle. Table 3

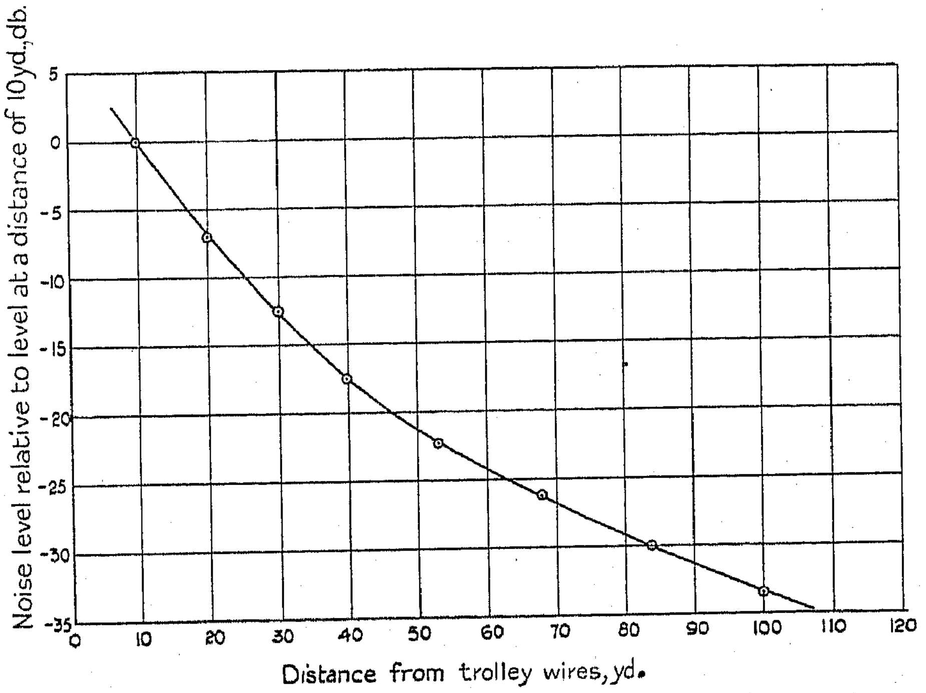

Fig. 6.—Variation of interference due to trolley-buses, with distance from track.

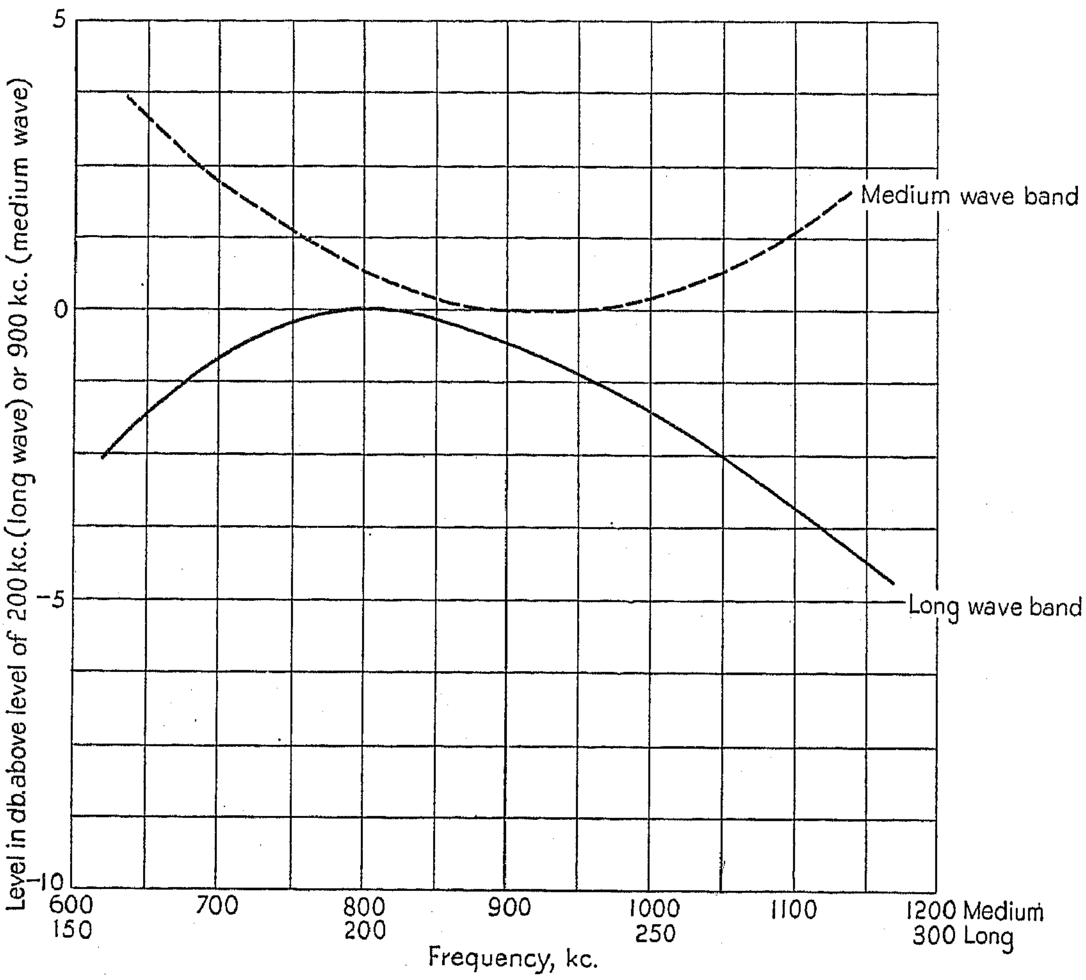

Fig. 7.—Variation of trolley-bus interference with frequency.

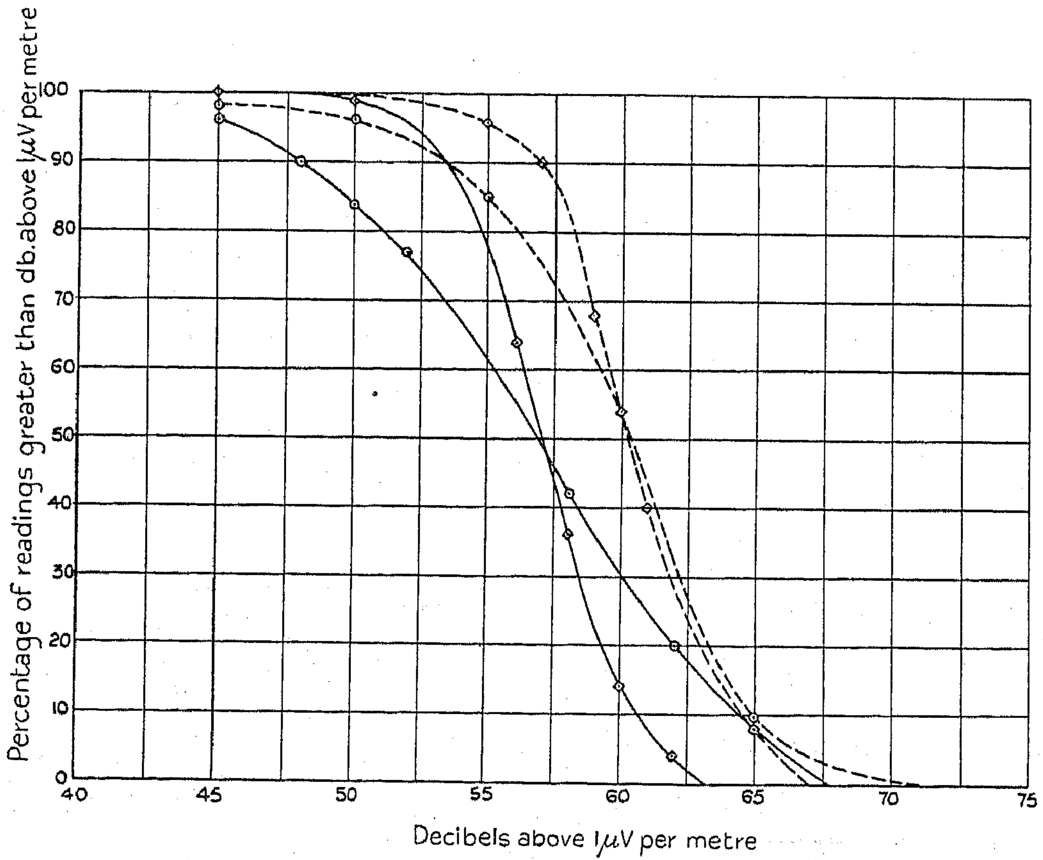

Fig. 8.—Intensity distribution curves of interference level due to trolley-buses. Observation point 10 yd. from track.

A{ ——◇——◇—— 200 kc.
— — ◇ — — ◇ — — 900 kc. } A. Single bus on a given route.

B{ ——○——○—— 200 kc.
— — ○ — — ○ — 900 kc. } B. Normal service.

shows values observed under different conditions for the various sources of interference.

It is observed that, although important where clinch ears are used and for severe conditions such as turning circles, section points, crossings, scaled or ice-coated contact wires, etc., the collector noise has been considerably reduced by the use of lubrication and grooved wires, skid collectors, spring holding devices, and carbon brushes, so that the more frequent and severe interference is due to the operation of contactors and their relay circuits. Motor noise is usually important in special and less frequent circumstances. The noise mainly associated with trolley-buses is therefore of the "click" type, which is not intolerable unless repeated with sufficient frequency. The interference therefore depends upon the traffic density, but it is not usually possible to take this into account until the number of audible "clicks" has been considerably reduced, either by distance (which, for this purpose, should exceed about $\frac{1}{4}$ mile) or by the application of suppressors.

Table 3

LEVEL OF INTERFERENCE FROM VARIOUS ITEMS OF A TROLLEY-BUS EQUIPMENT, IN DB. ABOVE 1 μV PER METRE

Frequency (kc.)	Type of braking	Collectors	Contactors	Main contactors	Main motor	Brake motor	Lighting generator	Distance (yd.)	
300	Rheostatic	73*	73	63	51	43		5	Company A, test track at Town A, line in oxidized and scaled condition
1 200	Rheostatic	75*	71		55	47		5	
300	Regenerative	77*	73		55	43		5	
1 200	Regenerative	73*	74		55	60		5	
1 000	Rheostatic	62†	53		23	5		Under lines	Town B
		63†	54		42	30		3 ft. from bus	
900	Regenerative	29+ } 5 yd.	73		46			Under lines	Town C
200		35+	82		54			12 yd. from bus	
900	Regenerative		103		48			12 yd. from bus	Town D
200			75		72			12 yd. from bus	
900	Regenerative and rheostatic							Under lines 12 yd. from bus	Town E
200									
900	Regenerative and rheostatic		84		53			do.	
200			78		50				
1 000	Regenerative and rheostatic		74	< 36	38			do.	Town F
200			72	< 36	40				
1 000			76		42	12		do.	Town G
200			72		45	24			
1 000			86	Negligible	18	25	16	do.	H
200			84	< 24	54	45	42	do.	Depot

* Wheel collectors.

† Clinch ears and skid collectors.

In order to give an impression of the relative frequency of disturbances of different intensity, Fig. 8 shows an analysis of the distribution of clicks of varying intensity for the representative conditions of a single vehicle on a given route and also of fairly frequent services.

(ii) High-voltage overhead transmission lines.

The interference arising from high-voltage overhead electric power transmission lines has not been brought so prominently to the notice of the general public as that set up by other agencies, since the greater portion of such lines lie in open country remote from residences. Interference nevertheless arises from such lines and may be conveyed by them from the point of origin to considerable distances and then radiated. In cases where high-voltage lines pass within 1–2 miles of commercial radio receiving stations working on long-distance services the

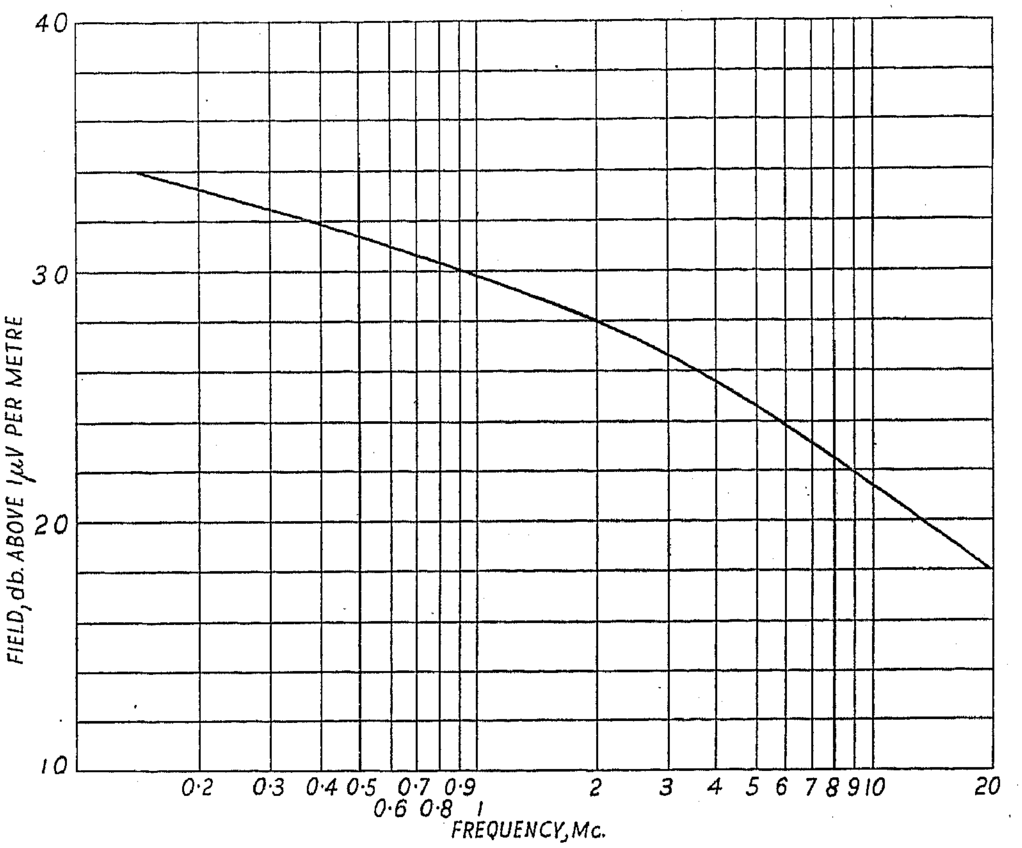

Fig. 9.—Noise field beneath 132-kV power line.

interference from the lines may be sufficiently serious to prevent the operation of the stations. To some extent this is due to the fact that such commercial stations normally work with a very low value of received signal and are usually placed in positions where electrical interference from other sources is absent.

The interference referred to is generated by the lines or by associated high-voltage switchgear, and can generally be traced to one of the following: arcing at contacts of switches, corona on lines, defective insulators causing corona or arcing.

Fig. 9 shows the frequency distribution over the range 150 kc.–20 Mc. of interfering energy immediately beneath a 132-kV power line at a point where a crackling noise could be heard from an insulator. Although there was no visible indication, even by night, of sparking or corona, continuous crackles were audible 30–40 yd. from the pylon. A series of random square-topped impulses would have given a spectrum in which amplitude was inversely proportional to frequency, but the properties of the line would tend to modify this distribution. It will be noticed, however, that the curve closely approaches this condition at the higher frequencies.

The observed attenuation of the interference on a number of selected frequencies in a direction normal to the power line is indicated in Fig. 10, and suggests that the rate of attenuation per mile decreases with distance. A similar result was obtained in observations of attenuation along the line. Fig. 11 gives the value of disturbing field, at a single frequency of 850 kc., beneath the line at various distances from the source of interference. The rate of attenuation decreases with distance and is much less in this direction than in directions normal to the line. Since the interference is radiated not only directly from the discharging items but also from the line which propagates the interference, the interfering field at any point in the neighbourhood of the system is made up of these component sources of radiation.

The strength of interference from new and clean insulators is of a much lower order of magnitude than that from insulators which have been subjected to weathering. For this reason the results of works tests at normal working voltage on insulators give little indication of the behaviour in service of a line equipped with such insulators.

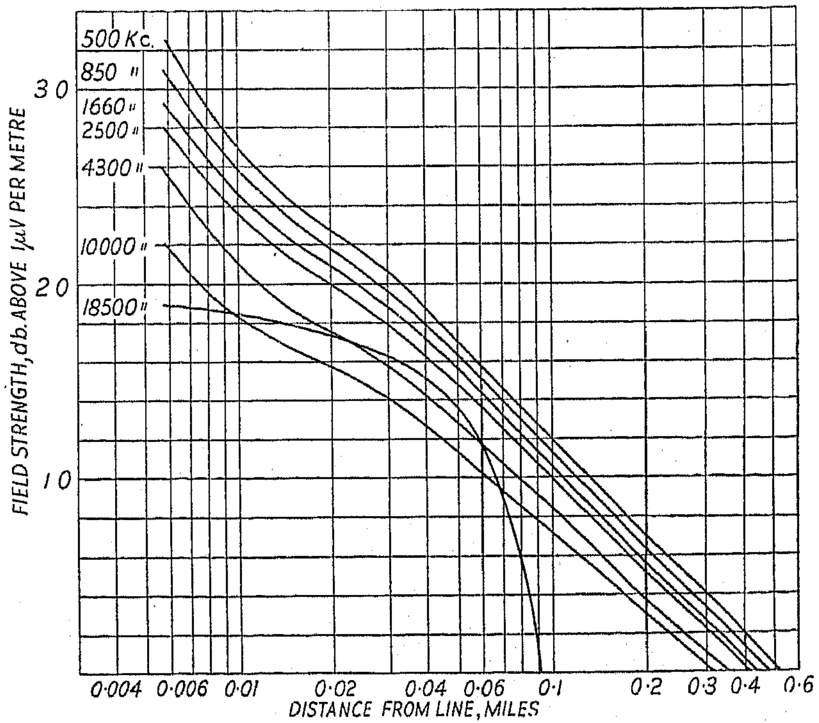

Fig. 10.—Attenuation curves of noise field in a direction normal to 132-kV power line.

In regard to the works testing of insulators, observations taken at the works indicated that the strength of the interference in the short wave band tends to increase with increase of frequency. The values of interference, in decibels relative to 1 microvolt per metre, from a 3-unit string of insulators tested at 63 000 volts, measured at a distance of 20 yd. from the source of disturbance, were as follows: at a wavelength of 15 m., 14 db.; at 20 m., 14 db.; at 30 m., 12 db.; at 40 m., 12 db.; at 60 m., 14 db.; at 80 m., 10 db.

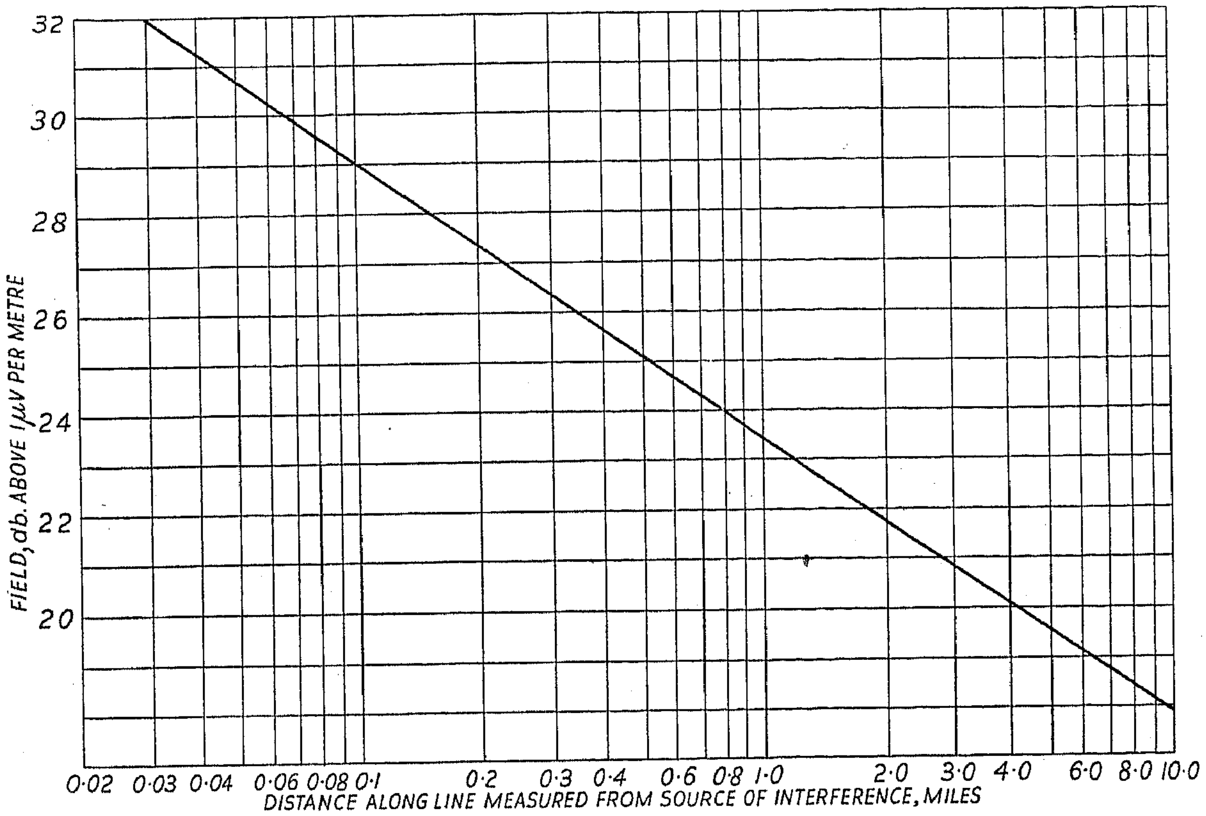

Fig. 11.—Field strength of noise beneath 132-kV power line on 850 kc.

For lines equipped with similar types of suspension and cap-and-pin insulators the strength of the interference appears to be independent of the line voltage of the system. The effect of weather conditions on the interference from a line is exemplified by Table 4, which gives the values measured at a distance of 15 yd. from a 132-kV line under wet and dry conditions respectively.

Table 4

Wavelength (m.)	Interference level (db. above 1 μV per metre)	
	Dry condition	Wet condition
15	-16	2
20	-12	8
200	16	42
350	20	46
1 550	28	56
2 000	30	58

In Germany, insulators are tested in the factory by observing the h.f. interfering current which flows when the insulator, under an appropriate applied voltage, is short-circuited by a condenser. This method depends upon the assumption that the impedance of the overhead line is small compared with the effective h.f. impedance of the insulator when causing interference in service.

(iii) High-frequency apparatus.

An obvious source of interference is apparatus which generates and utilizes h.f. currents, such as h.f. medical and surgical equipment (e.g. electrotherapy and diathermy), and h.f. furnaces. The radiation becomes greater as the working wavelength of the apparatus is reduced and as the dimensions of unscreened portions of the circuits approach the order of magnitude of the wavelength.

The most common source of interference of this class is electromedical apparatus, which includes equipments of moderate or high power used under professional supervision for therapy and diathermy and also the smaller portable equipment known as " violet-ray apparatus." The fundamental difference between electrotherapy and diathermy is that in the former there is no conductive connection between the apparatus and the patient, energy being conveyed through the capacitance of air-gaps, while in the latter there is a conductive connection between the apparatus and the patient. In both types of equipment h.f. currents are generated, either in the form of damped waves by a simple type of spark transmitter circuit or by the use of thermionic valves energizing an oscillatory circuit. The output h.f. voltage in each case is stepped up to a high value by means of a transformer and then applied to the patient by means of either one or two electrodes.

When a single electrode is employed, as in the case of the small portable type of violet-ray apparatus, the secondary currents return via earth capacitances, and intense radiation at high frequencies occurs. With electrotherapy and diathermy apparatus, however, two electrodes are employed and the secondary circuit consists of a smaller loop than in the single-electrode case. In consequence the field directly radiated may not be quite so severe. It is found, however, that the field around the apparatus, patient, and connecting leads will still be sufficiently intense to interfere seriously with radio reception.

The mains-borne interference from both types of plant will always be serious and may be propagated along the supply mains for several hundreds of yards. As such apparatus is frequently installed and used in residential districts, it may thus become a source of interference to radio listeners over a wide area.

In general, interference set up by this type of plant will be more or less tunable and will have a maximum at the frequency of generation. A large proportion of the equipments at present in use operate on the long- and medium-wave broadcast bands. Some of the latest types of apparatus which are coming into general use in this country are capable, however, of interfering very seriously with the reception of television on ultra-short wavelengths, and the range of this interference may be as much as 5 miles. Cases have been reported of interference in this country on short wavelengths between 10 and 100 m. From the fact that this interference had a pronounced 60-cycle modulation, it has been assumed that it was caused by electromedical apparatus in America.

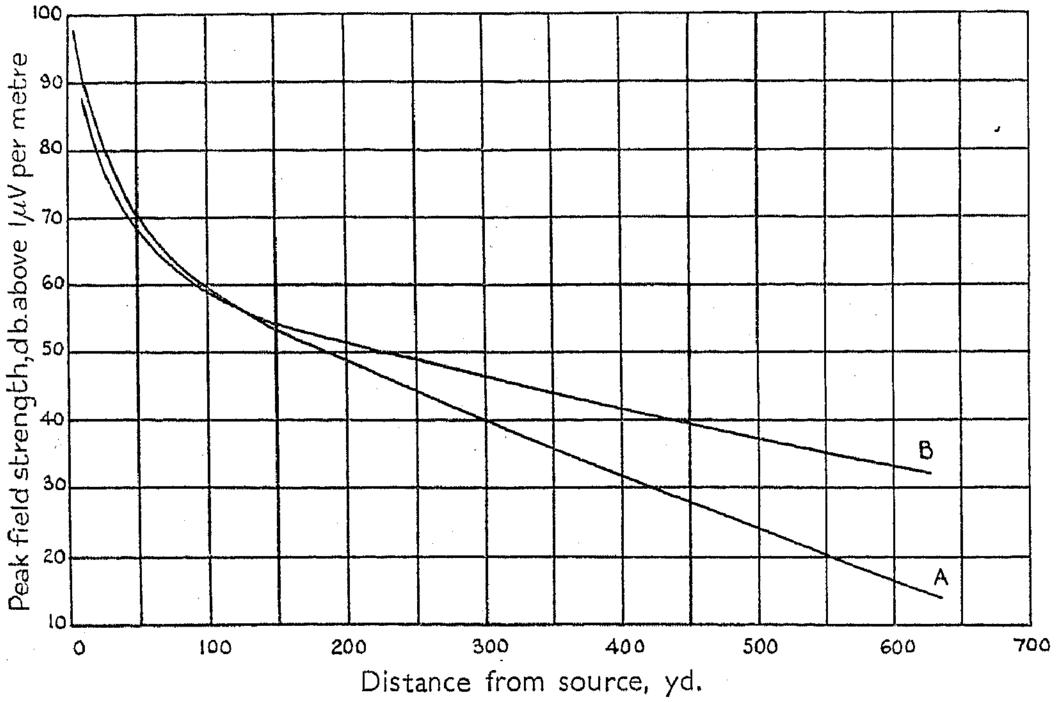

Fig. 12.—Variation, with distance from source, of radiation field given by 30-metre 0·5-kW valve therapy machine. Output current, 2 A.

A. 2nd harmonic (24 Mc.).
B. 4th harmonic (48 Mc.).

Particulars of the levels of interference from therapy and diathermy plant on long and medium waves are given in Section (3) (*a*) (vi). Tests have been made by the E.R.A. as regards ultra-short waves on the following therapy and diathermy plant: (1) 0·5-kV 30-m. valve generator, (2) 0·3-kV valve generator (30 and 6 m.), (3) spark machine (15 to 6 m.), (4) spark machine (300 m.). An artificial load consisting of a wooden block or a dish of solution was used, and the field strength measured by the short-wave receiver mounted in a van.

The valve machines emitted radiation only at the fundamental and harmonic frequencies. Fig. 12 shows the variation with distance of the field due to the harmonics of Machine (1). The associated noise was small owing to the use of rectified and smoothed H.T. supply to the valves, the a.c. supply leads being also fitted with h.f. chokes. Machine (2) was supplied with raw alternating current and consequently the field produced was heavily modulated; metal panels were fitted to the machine but were not bonded and had apparently little screening effect. The removal of the leads to the patient reduced the radiation by about 10 db. at a fundamental frequency of 42·5 Mc. The variation of the field of fundamental frequency with distance is shown in Curve A, Fig. 13, while Curves B, C, and D refer to the 3rd, 4th, and 5th harmonics respectively, when a lower fundamental frequency (7·6 Mc.) was used. The band width of the radiation on the fundamental and harmonics of both valve machines varied from 500 to 150 kc. when referred to a level 40 db. below the maximum, and measured with the short-wave measuring set already described. The field from Machine (3) at 37·4 Mc. is shown in Fig. 14. With this machine the patient forms a part of the output tuned circuit. As opposed to valve machines, spark machines emit radiation over a wide band of frequencies. For Machines (3) and (4), Fig. 15 shows the radiation to be distributed over the band from 20 to 50 Mc.

It is obvious from these results that h.f. therapy machines of such types are capable of causing serious interference to television reception over a considerable area, on both the fundamental frequency and the harmonics of valve machines and over the whole band in the case of spark machines.

(iv) Electric lifts.

The equipment of a lift comprises interfering items in the form of the driving motor, giving continuous noises, and of the contactor relay and control circuits, which give noises of the "click" type. This equipment is often installed in buildings (and ships) where radio receivers

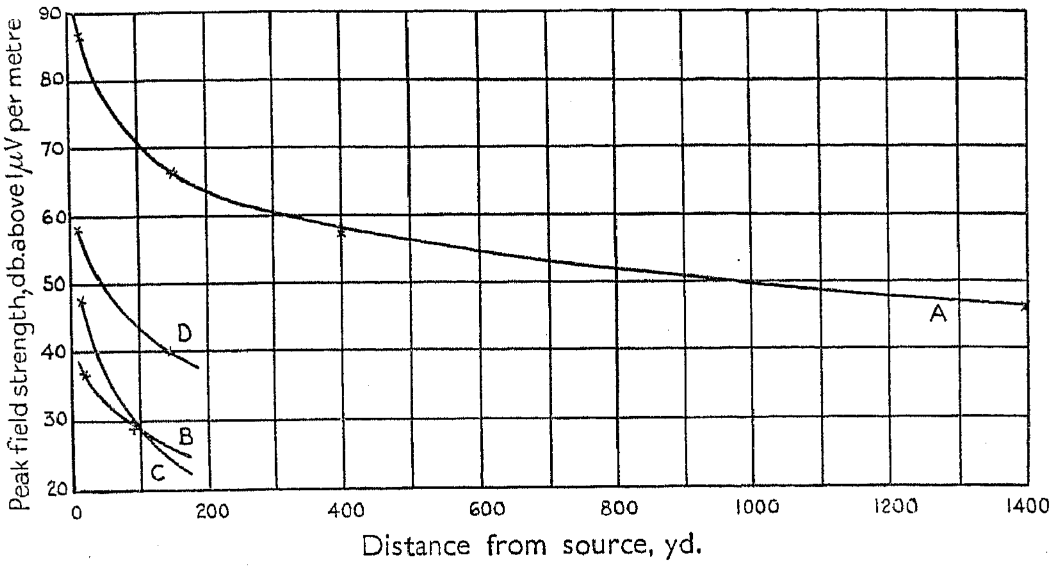

Fig. 13.—Variation, with distance from source, of radiation field given by 30–6 metre 0·3-kW valve therapy machine. Output current, 4 A.

A. Fundamental of 42·5 Mc.
B. 3rd harmonic of 7·6-Mc. fundamental.
C. 4th harmonic of 7·6-Mc. fundamental.
D. 5th harmonic of 7·6-Mc. fundamental.

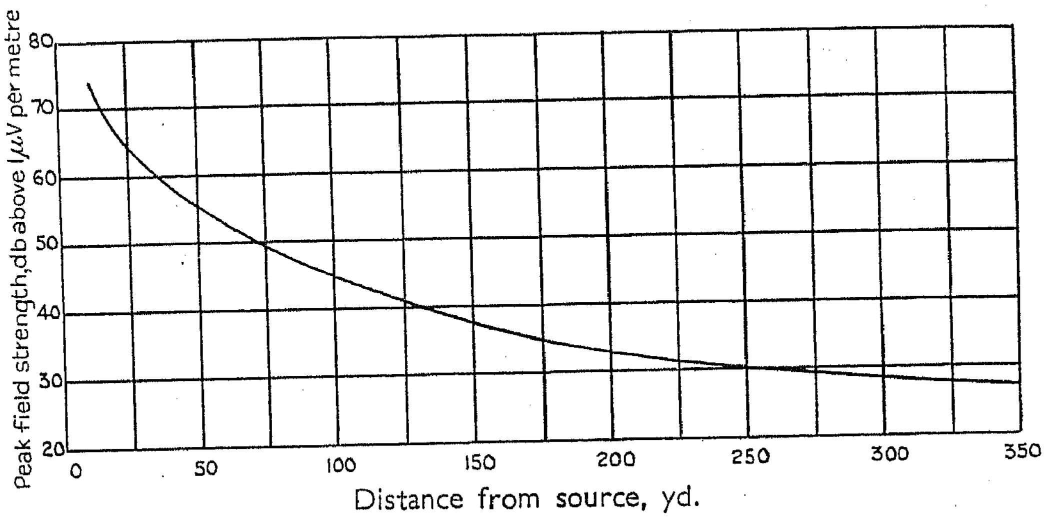

Fig. 14.—Variation, with distance from source, of radiation field given by 6–15 metre spark therapy machine. Frequency 37·4 Mc., output current 2 A.

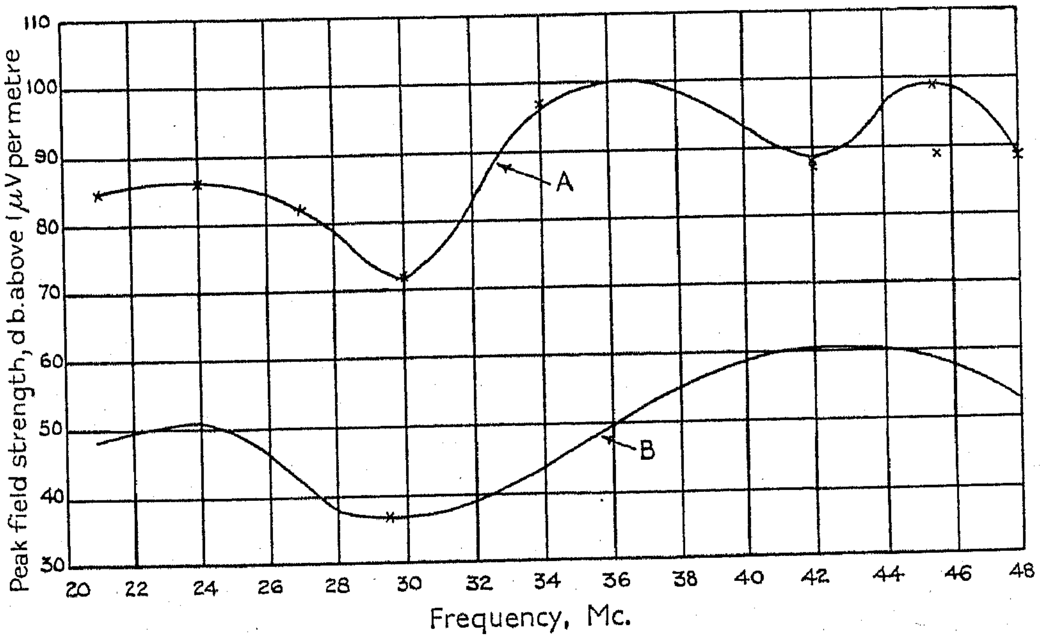

Fig. 15.—Frequency variation of intensity of radiation for spark machines.

A. 30–6 metre spark machine, output current tuned to 37 Mc.
B. 300-metre spark machine.

operate in close proximity. The radiation is further increased by the fact that the interfering sources are connected or may be coupled to the trailing cable which connects the panel in the lift car to the main equipment, and to the wiring between the gate interlocks, switches, etc., and the control gear and supply. Whereas the latter wiring is often run in steel conduits the trailing flexible cable is suspended between the car and the halfway box, normally without metallic protection. A schematic diagram is shown in Fig. 16.

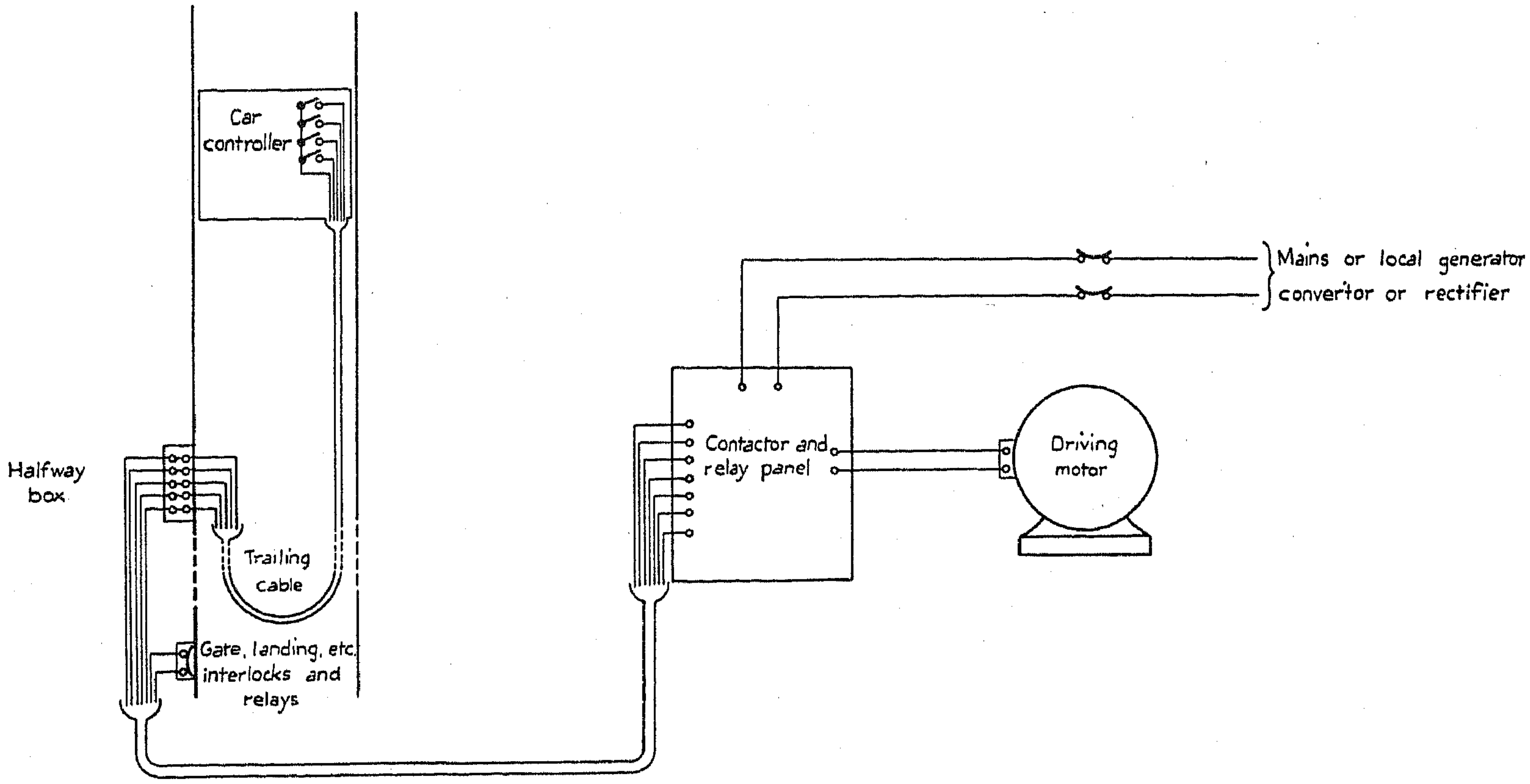
Car controller
Mains or local generator convertor or rectifier
Halfway box
Contactor and relay panel
Driving motor
Trailing cable
Gate, landing, etc. interlocks and relays

Fig. 16.—Simplified circuit scheme for lifts.

Interference may thus reach the listener's receiver by way of the electric mains, if on the same system as the normal domestic services, or by direct radiation from the motor and main control gear, from wiring in the shaft or from the trailing cable acting as a radiating aerial. In addition, it sometimes occurs that the lift shaft is wholly or partly enclosed by a metal network, which may be inadequately earthed from an h.f. point of view. H.F. voltages may then be developed along the shaft which, although small, can give serious interference since the coupling with the listener's aerial may be high. Tests made on a lift partially enclosed by wire netting yielded voltages of the order of 1 mV developed therein between the halfway level and earth. This value could be considerably reduced by efficient earthing of metalwork.

The radiated motor interference is less than that due to the control circuits, for well-maintained motors. The "motor noise" is negligible for lifts operated on motor-generator systems, while for d.c. motors with direct supply the noise is generally at least 10 db. below the level of other disturbances, except where faulty commutation or other defects cause a high interference value. Auxiliary motors for automatic gate-closing must also be considered, but follow the same principles as for other small motors. Control circuits energized with alternating current, as in some modern types of lift, cause little disturbance. The main interference is due to the circuits controlling the operating coils of the contactors and switches of the driving motor and also the brake-magnet coil. These circuits ascend the shaft and are operated by hand or automatically in the car or at the gates. They are most troublesome when supplied direct from the mains, being then usually highly inductive. In some older types of lift, the manual control circuits carry the holding, closing, or tripping currents and are then a source of severe interference. In general, all the circuits for remote control, particularly those which enter the trailing cable, form systems which give an h.f. radiation, determined by the inductance of the coils and the self-capacitance of the circuit involved.

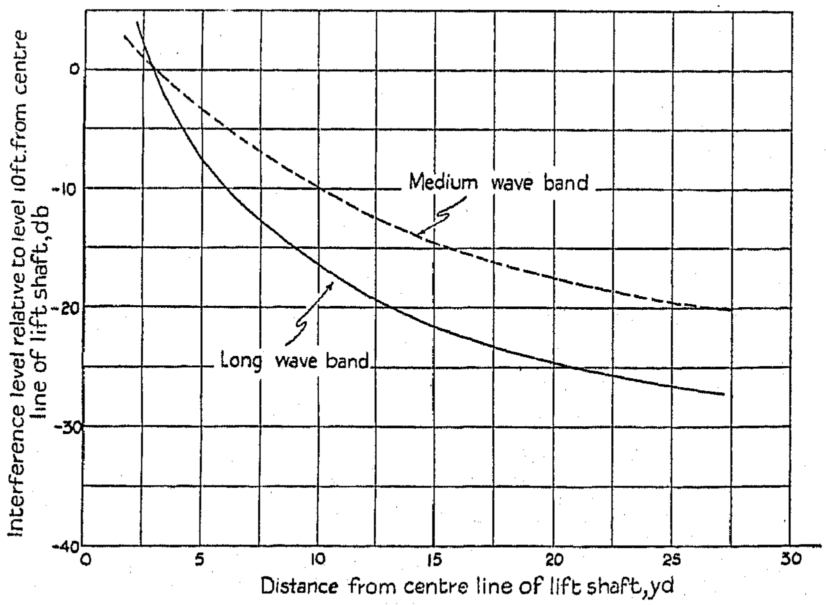

Fig. 17.—Variation of interference with distance from centre line of lift shaft.

The brake-magnet coil is also a radiating source, although its current is confined to the main installation. The other heavy-current circuits such as those associated with the motor do not usually radiate much interference on closing or interruption. Tests in which these circuits were manually operated showed that their effect might be neglected in comparison with that of the control circuits, at least at distances greater than about 10–20 ft. from the main installation.

Table 5

UNSUPPRESSED LEVEL OF RADIATED INTERFERENCE FROM ELECTRIC LIFTS

Lift No.	Type of car control	Control-circuit supply	Distance of test position from trailing cables (ft.)	Average interference level (db. above 1 μV per metre)			
				900 kc.	1 000 kc.	190 kc.	180 kc.
1	Hand-driven controller, direct control	D.C.	10	77–65	—	—	62–53
2	Push-button remote control . .	D.C.	10	45	—	—	30
3	Push-button remote control . .	A.C. supply, a.c.–d.c. motor-generator	10	28	—	—	40
4	Hand-driven controller	D.C.	10	—	30	39	—
5	Push button	D.C.	10	—	37	34	—
6	Push button	A.C. supply rectified by mercury - arc rectifier	10	—	44	52	—

Table 5 shows values of the field due to some representative types of lift. Some figures indicating the rapidity of attenuation with distance from the lift are given in Fig. 17. The variation of interference with frequency is not marked.

(v) Telecommunication apparatus.

Telecommunication apparatus can, in certain circumstances, give rise to radio interference. In the case of automatic telephone plant the interference may be caused by plant at the exchange or at the subscriber's premises. In the exchange the numerous relays and selectors constitute a source of interference, although this is greatly reduced by the present practice of fitting spark-quench circuits to apparatus for the preservation of contacts. When underground distribution is employed, little or no interference occurs except in the immediate vicinity of the exchange building and this can generally be relieved by re-positioning the listener's aerial. Where overhead distribution is used the interference can be troublesome, and in view of the multitude of switching contacts concerned it is usually more economical to filter all circuits leaving the exchange building than to apply filters to all sources of interference within the building. Table 6 gives the interference from a small exchange measured at a point 15 yd. from the exchange building.

Table 6

Frequency (kc.)	Signal (db. above 1 μV per metre)	Noise (db. above 1 μV per metre)		Signal/noise ratio (db.)	
		Before suppression	After suppression	Before suppression	After suppression
877 (London Regional)	34	42	-8	-8	$+42$
200 (Droitwich)	63	23	$+2$	$+40$	$+61$

The fitting of filters at the exchange building does not eliminate the interference produced by the subscriber's own dial. This interference, however, is only likely to affect receivers in the immediate vicinity of the subscriber's instrument and line. It can readily be eliminated where troublesome by fitting a dial suppressor.

Machine telegraph apparatus is also a potential source of interference, but the number of cases where such interference is troublesome is relatively small as the apparatus and circuits are not normally present in areas where broadcast reception is prevalent. Moreover, apart from direct radiation, which is of comparatively short range, the trouble is usually caused by radiation from overhead lines, which are rarely employed in areas where machine telegraphs are installed. The chief difficulty arises with such equipment when it is installed inside a radio-telegraph receiving station handling long-range traffic. Here the interference can be sufficiently great in relation to the average signals to embarrass the service.

(vi) Ignition systems.

Although installations such as oil-burning furnaces comprise ignition systems which may give rise to appreciable interference, the motor-car is so eminently the most widespread example that it deserves special study. Both the low-voltage equipment and the high-voltage circuits form radiating systems.

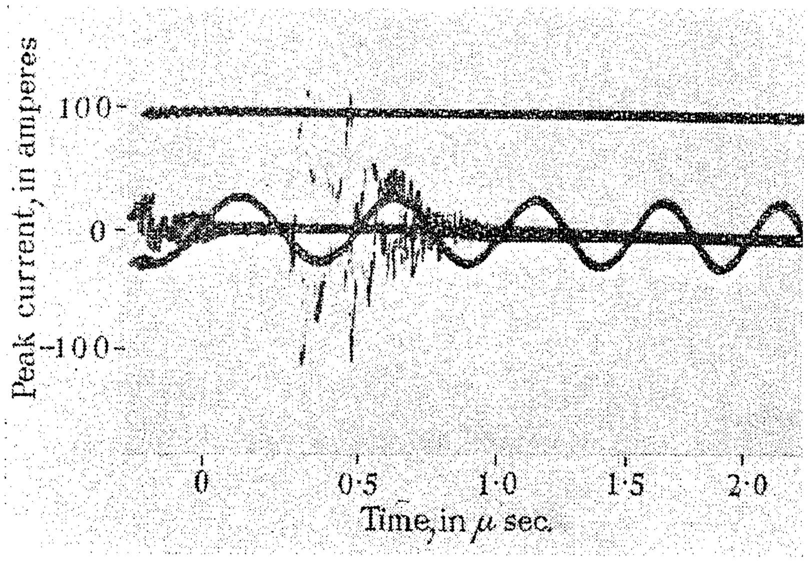

Fig. 18.—Oscillatory spark currents.

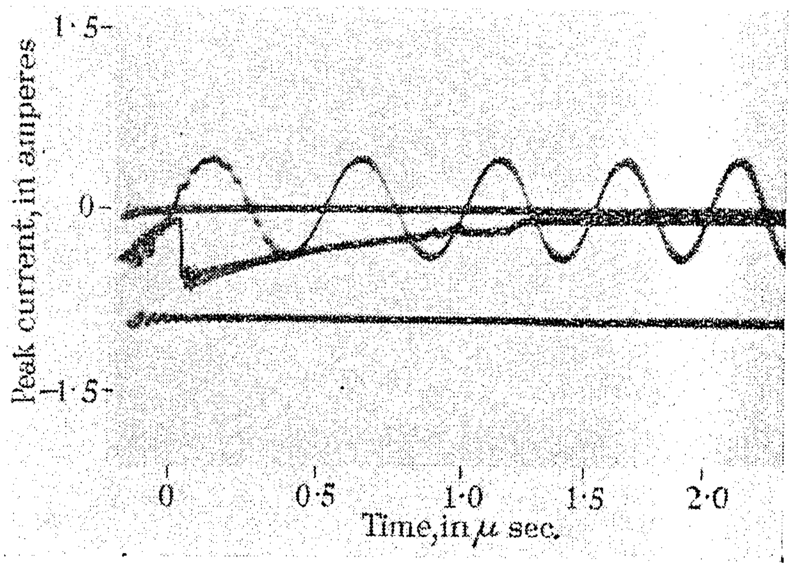

Fig. 19.—Non-oscillatory spark currents (10 000 Ω resistor in plug lead.)

In the latter, the circuit formed by the plug, plug lead, distributor, coil or magneto, and chassis, is excited by the discharge at the plug and the spark at the distributor (if a jump-spark distributor is used), giving rise to a train of very high-frequency damped waves repeated at each discharge, as shown in Fig. 18.* The current may attain instantaneous values of the order of 100 amp. If resistance is present, as when a suppressor is inserted at the plug (see Fig. 19*), the current transient becomes non-oscillatory and of very steep wave-front, but the crest value is very much reduced. These rapid transients are succeeded by oscillations of longer period. The radiation extends over a wide frequency-range. The lower frequencies, covering the lower-frequency broadcast bands, appear to be determined by the inductive discharge of the coil or magneto. The effective inductance of the latter with the self-capacitance of the circuit gives an oscillation spectrum, similar to a circuit with distributed constants subject to shock excitation. From the interference point of view, the higher frequencies, particularly the range 10–80 Mc., are more important, since the radiation is more intense and the services envisaged for this band are of lower field-strength. At these frequencies it appears that the alternating components of the initial disturbance are radiated from the distributor and plug leads, which act as aerials. The audio-frequency response thereby produced in a receiver is a series of impulses of which the time of passage is of the order of 3 millisec., which is much less than the interval between successive impulses, this interval being that between successive sparks.

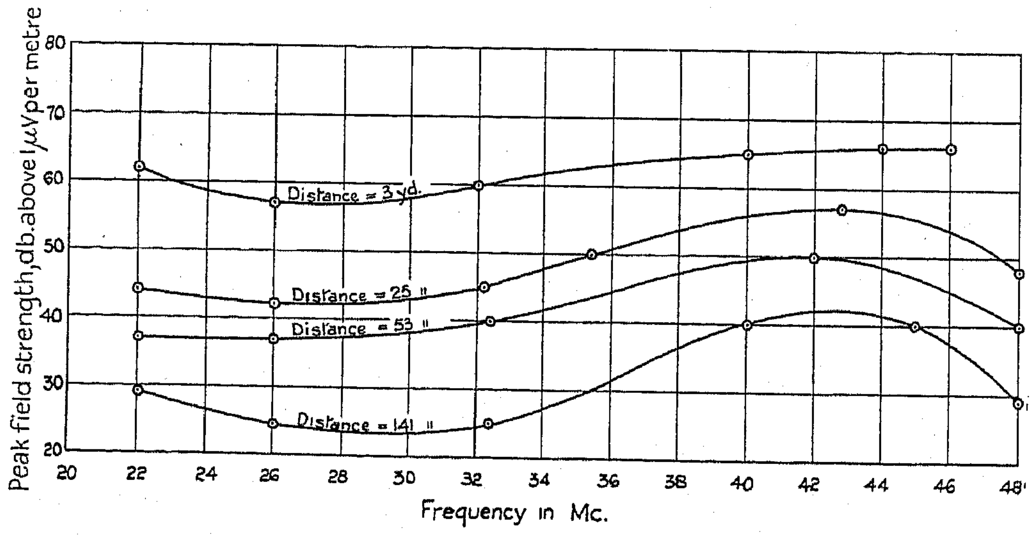

Fig. 20.—Frequency variation of average peak field-strength of interference from automobiles. Values obtained on Western Avenue.

Fig. 20 shows the variation of interference with frequency observed near an arterial road, and Fig. 22 shows the attenuation with distance. The order of levels observed and their relatively slow attenuation with distance indicate that motor-cars must be regarded as important potential interfering sources. Fig. 21 shows the results of tests on a number of automobiles at a standard distance of 10 yd. (ground level) and a standard speed of 30 m.p.h. For a level of 50 μV per metre (provisionally adopted) no correction is usually required at 13 Mc. and only about 15 db. at 45 Mc. All the systems considered later are capable of giving suppression of this order, so that automobiles suppressed for receivers mounted thereon will also be suppressed as regards external receivers. It is common in modern cars to mount the coil on the bodywork, and this practice gives a long distributor lead. Recent E.R.A. tests have shown that if the coil is suitably mounted on the engine so as to reduce the dimensions of the radiating circuit and confine it to the engine, the interference levels just quoted may be reduced by 8–21 db. By redesign of the layout on this and similar bases it is possible in some cases to avoid the necessity for the use of actual suppressors.

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* Oscillograms obtained by W. NETHERCOT, of the E.R.A., on discharges in air at normal temperatures and pressures.

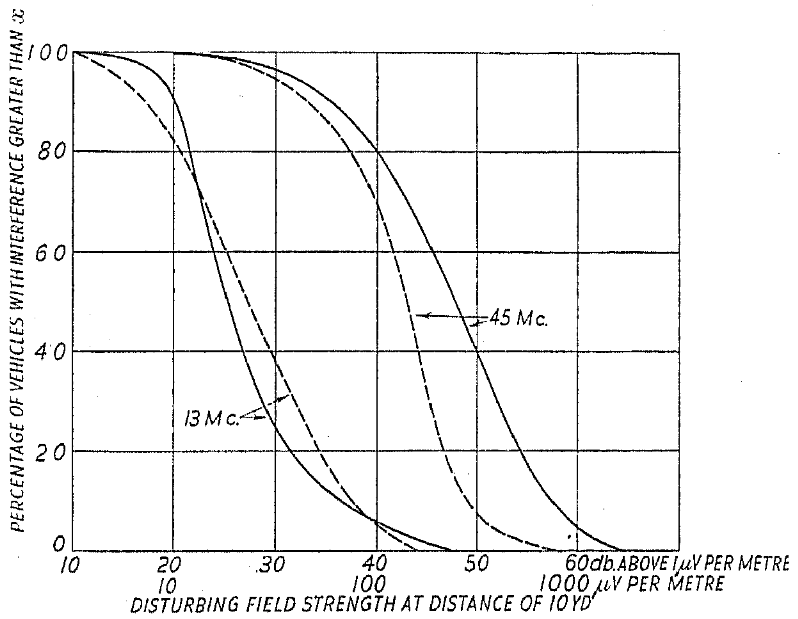

Fig. 21.—Interference due to automobiles.

———— Private cars.
— — — — Commercial vehicles.

The lower-frequency radiation from the high-voltage circuits and the radiation from the low-voltage equipment is mainly of importance with respect to receivers in the vehicle. As regards the low-voltage equipment, it may be mentioned that auxiliaries such as voltage regulators and windscreen wipers are frequently the most severe sources of disturbance. The disturbance is radiated from the car wiring as a whole and, though intense in its immediate neighbourhood, is rapidly attenuated with distance.

(b) Re-radiated Interference

(i) Domestic items.

The increasing use of electrical appliances for domestic purposes has brought the question of interference prominently to the notice of the large mass of the general public, as the source of interference is in their own homes or in adjacent premises and the disturbance from this class of plant is concerned almost solely with reception of broadcast programmes. The sources of interference can be divided broadly into two groups, namely motors and switching devices. Some types of equipment contain both sources of interference, e.g. electric refrigerators using commutator motors. The normal path of interference is from the interfering item via the supply mains to the vicinity of the receiving aerial, whence it is radiated from unscreened portions of the wiring and picked up by the aerial-earth circuit of the receiver. The amount of interference created by such devices will depend on the interference voltage set up by the appliance, the internal impedance of the appliance to the h.f. disturbance, and the attenuation in the path between the device and the receiving antenna. The interfering device can be regarded as a radio-frequency generator with a definite internal impedance supplying energy to the mains.

Owing to the fact that some countries already had legislation requiring the suppression of interference in such devices, it became necessary to attempt to secure

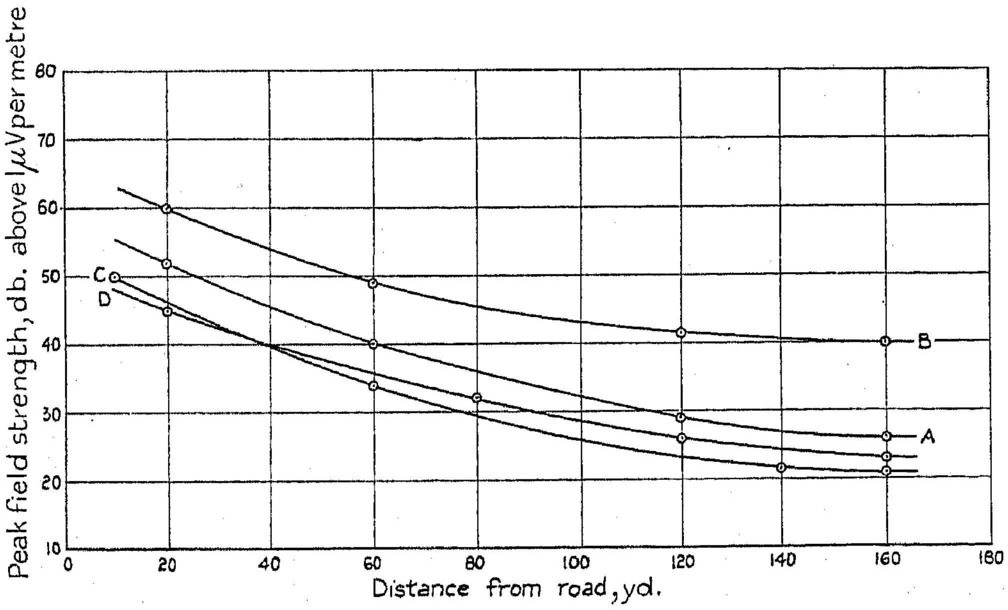

Fig. 22.—Variation of average peak field-strength of interference from automobiles, with distance from road. Values obtained on Western Avenue.

A. Frequency = 48 Mc.
B. Frequency = 45·5 Mc.
C. Frequency = 26·0 Mc.
D. Frequency = 21·0 Mc.

international agreement as to the value of noise voltage which could be tolerated in such plant. With this object in view the I.E.C. appointed in 1934 a Special International Committee on Electrical Interference (C.I.S.P.R.) to study the subject. In connection with the work of this Committee a number of measurements have been made in this country of the interference voltage generated by domestic and other low-power appliances. Measurements have also been made of the effective heights of listeners' aerials, the coupling between the receiving aerial and the mains, and the impedance of the latter.

For the purpose of the measurements the effective height of an aerial was defined as the ratio between the e.m.f. measured on open circuit across the aerial and earth terminals of the receiver in its usual situation and the value of e.m.f. which would be obtained with an aerial of an effective height of 1 m. placed in a free position outside the listener's premises.

The coupling between the receiving aerial and the interfering item was determined by the application of a sinusoidal h.f. voltage of known value symmetrically and asymmetrically between the mains terminals of the interfering apparatus, and between these terminals and earth, by means of a portable calibrated generator substituted for the machine or apparatus, and the measurement of the open-circuit e.m.f. induced between the aerial and earth terminals of the receiver. It was also decided to adopt a value of the order of 1 volt for the voltage applied to the mains, in order to reduce errors which might be caused by the presence of other noises.

The British measurements were carried out by the G.P.O. and were made at the houses of 214 listeners. Of the premises visited, 40 % were those of members of the staff of the Post Office Engineering Department at Rugby, Baldock, St. Albans, and N.W. London, 5 % were those of officers of the B.B.C., 20 % were those of members of the general public who had complained of electrical interference with broadcast reception, and the remainder were those of members of the general public who agreed to measurements being made on being told of the nature of the experiments. Such members were selected at random from districts which were as far as possible uniformly distributed over London. Post Office field-strength measuring sets of the type shown in Fig. 2 were utilized for measuring the voltage developed between the earth and aerial terminals at the listener's premises by the insertion of a high resistance in series with the normal aerial input connection of the measuring set to increase the input impedance of the set; the e.m.f. to be measured being applied between the remote end of the resistance and the casing of the set. Sets modified in this manner have been found convenient for the measurement of the effective height of receiving aerials, the set being used first as a voltmeter for measuring the voltage induced in the antenna by a broadcast transmission and then as a field-strength measuring set. The ratio of the signal thus measured, in microvolts (open circuit), to the field strength of the radiation in free space near the antenna, in microvolts per metre, gives the equivalent effective height in metres. The measurements were made on frequencies of 200, 876, and 1 150 kc.

A summary of the data obtained regarding effective heights of aerials is given in Fig. 23. The probable accuracy of the measurements is estimated as ± 2 db. (approximately ± 20 %). It was found during the tests that the proportion of listeners using outdoor aerials was 70 %. Analysis of data curves of the effective heights of the aerials of the listeners who had complained of interference disclosed that these curves did not differ appreciably from those of Fig. 23 for the total number of aerials. Separate curves were also made of the effective heights for indoor and outdoor aerials on the three frequencies. As a rough generalization, it can be stated that the effective heights of outdoor aerials were of the order of 3 times the effective heights of indoor aerials.

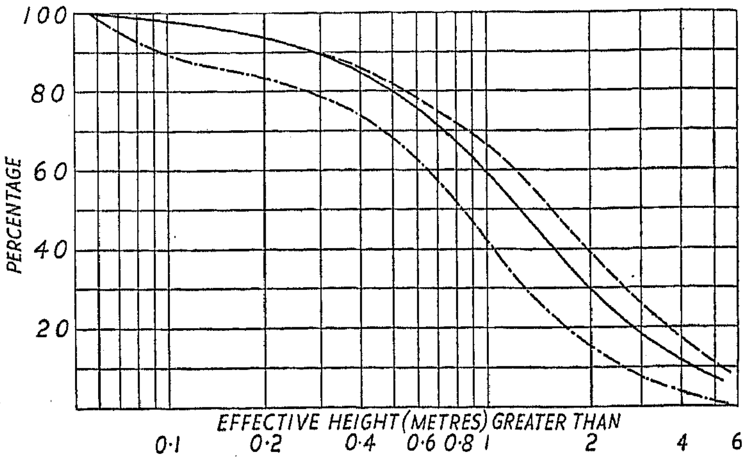

Fig. 23.—Frequency curves of the effective heights of aerials on various wavelengths.

——————— 261 m.
— — — — — — 342 m.
—·—·—·—·— 1 500 m.

The coupling between the mains and the aerial, expressed as attenuation for symmetrical and asymmetrical conditions, is given in Figs. 24 and 25. It is interesting to note that no relationship was apparent between effective height and coupling.

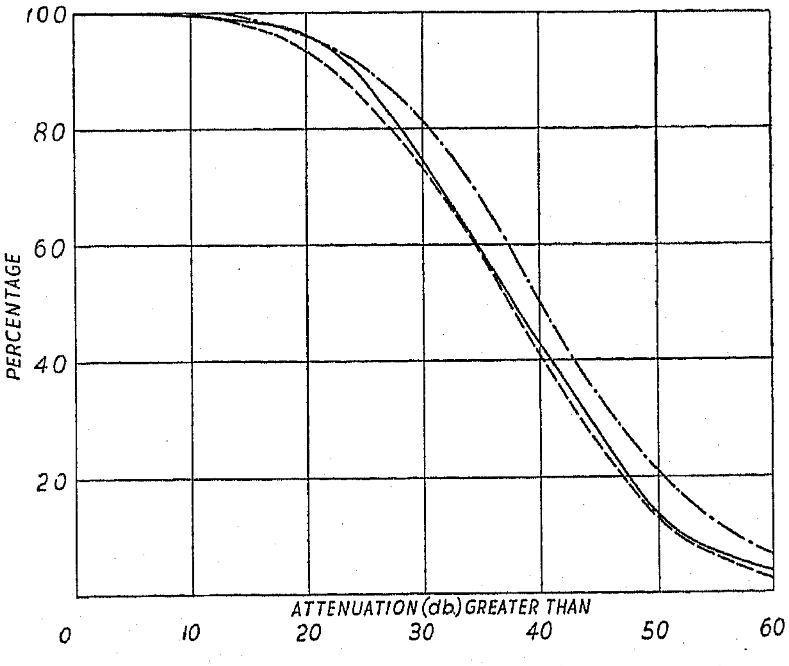

Fig. 24.—Transmission characteristics of domestic mains on various wavelengths for the symmetrical condition.

——————— 250 m.
— — — — — — 450 m.
—·—·—·—·— 1 600 m.

Interference can be propagated along the supply mains either as a difference of potential between the two conductors themselves or as a difference of potential between the mean potential of the conductors and earth. As previously mentioned, the former path is generally referred to as the symmetrical path and the latter as the asymmetrical path. Owing to mutual cancellation of effects the radiation due to currents flowing in the symmetrical path is generally much less than the radiation from currents flowing in the asymmetrical path. The impedance of the mains was measured for both the symmetrical and the asymmetrical condition by applying a sinusoidal h.f. voltage through a calibrated variable capacitance to the unknown impedance. The impedance could then be determined from a knowledge of the variable capacitance, the frequency and voltage of the oscillator, and the voltages across the variable capacitance and across the unknown impedance.

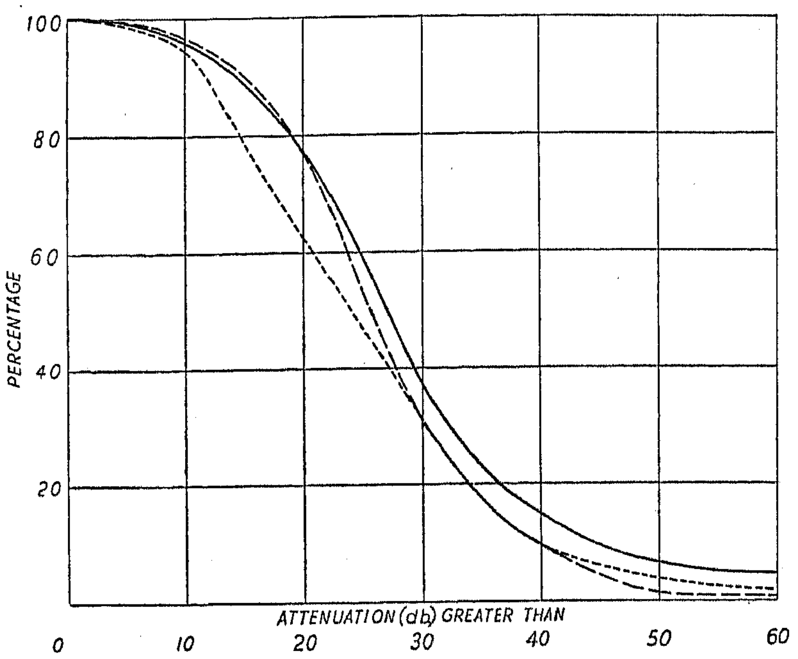

Fig. 25.—Transmission characteristics of domestic mains on various wavelengths for the asymmetrical condition.

———————— 250 m.
— — — — 450 m.
- - - - - - - - - 1 600 m.

The measurement of the interference from machines necessitates the use of a mains network to ensure that the appropriate voltage is selected for measurement and that other noises propagated along the mains are not included in the measurement. A suitable form of mains impedance network, shown in Fig. 26, provides for the impedance of the mains to be balanced with respect to earth. For measurement of the symmetrical component the measuring instrument is applied across the resistance network of 200 ohms shunted by 600 ohms, while for measurement of the asymmetrical component the 200-ohm resistance is short-circuited and the two 300-ohm branches connected in parallel between the mains and earth. Thus in each case the interference path is terminated by a resistance of 150 ohms. Recently it has been provisionally agreed to measure the asymmetrical component between the mid-point of the 200-ohm resistor and earth, avoiding the short-circuit between the mains terminals. The test results are usually the same with either method. The two chokes prevent any interference voltage present from the mains being introduced into the measuring set. In the U.S.A. a value of 600 ohms has been adopted for the equivalent mains impedance.

The impedance of the supply main was found to vary over wide limits. In general it can be said that the impedance will be between 30 and 300 ohms for the symmetrical condition and between 100 and 1 000 ohms for the asymmetrical condition, for frequencies of 700 to 1 200 kc., and from 15 to 150 ohms and from 30 to 200 ohms respectively for the two conditions at a frequency of 200 kc. The value of 150 ohms is arbitrary, but a standard is essential since the interfering voltage varies to some extent proportionately to the mains impedance.

With regard to the noise voltage from interfering items measurements have been made, at the time of writing, on 352 electrical appliances of various kinds, as follows:—

(*a*) *Domestic appliances* (*up to* 500 *watts*)	No.
Bells	6
Dusters (hand)	2
Fans	34
Floor polishers	5
Heating pads	1
Lamps	2
Mixer (juice extractor)	3
Gramophone motor	2
Refrigerators	17
Sewing machines	8
Vacuum cleaners	43
Violet-ray apparatus	4
Washing machines	3
Water heater	3

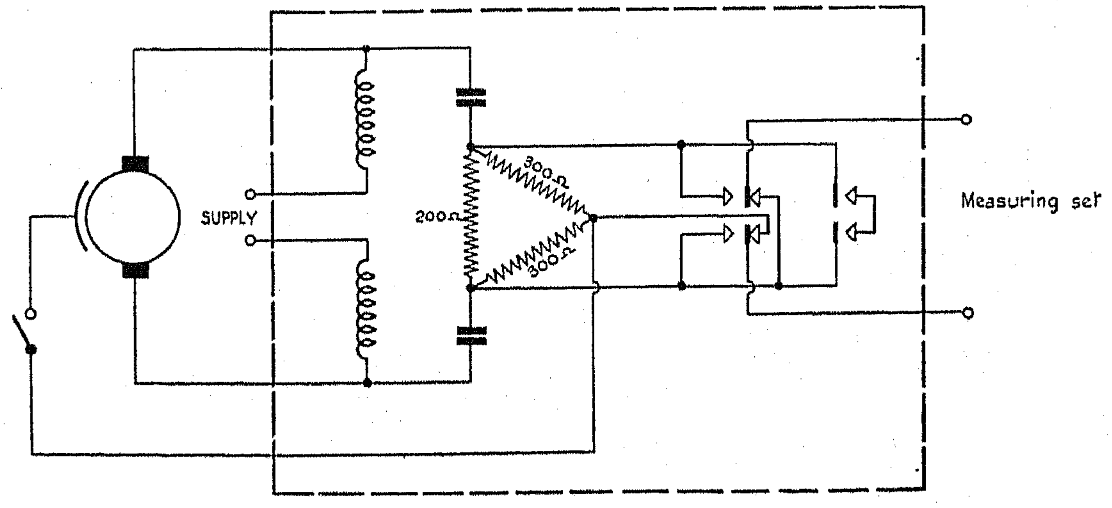

Fig. 26.—Artificial mains impedance network.

(*b*) *Non-domestic appliances* (*up to* 500 *watts*)	No.
Adding machines	3
Coffee grinders	2
Dental apparatus	12
Diathermy apparatus	2
Flashing signs	4
Furnace (electric)	1
Hair clippers	6
Hair cutters	1
Hair dryers	20
Meters (clock, etc.)	2
Motors	38
Neon signs	6
Rectifiers	1
Rotary convertors	4
Thermostats	3
Tools (portable electric)	85
Traffic signals	2
Turntable (motor display)	2
Valve rectifiers	2
Vibrators	6

Table 7

Category	Level of unsuppressed interference* (db. above 1 μV)							
	Maximum				Mean			
	At 190 kc.		At 1 000 kc.		At 190 kc.		At 1 000 kc.	
	S	A	S	A	S	A	S	A
Category A (*up to* 500 *W*)								
(i) Domestic appliances, frame earthed	100	103	92	95	70	63	67	61
(ii) Non-domestic, frame unearthed	124	90	96	81·5	72	50	70	50
(iii) Non-domestic, frame earthed	124	116	102	97	77	76	66	68
Category B (500 W to 10 kW)	122	115	121	116	88	82	81	75

* S = symmetrical, A = asymmetrical.

(*c*) *All types of machines*, 500 *W*–3 *kW*	No.
Dust-precipitation plant	1
Furnaces (electrical)	2
Motors	6
Ovens	3
Starting panel	1
Welders	4

The magnitude of the unsuppressed interference from these appliances was as shown in Table 7.

(ii) Electric lifts.

The interference from the sources, already enumerated in Section (2) (*a*) (iv), which may be comprised in a lift installation, appears also as interfering voltages at the supply terminals, which, as regards origin, type, and frequency spectrum, are similar to the interference fields already described. Table 8 shows figures obtained with some representative types of installation. When a mercury-arc rectifier or rotary convertor is used for local supply, the mains noise of the installation is that appropriate to such items, which are considered separately elsewhere.

The h.f. impedance of a lift installation is indeterminate for noise of the click type. For motor noise the values correspond to those for large motors. Tests on Lift (ii) (Table 8) gave 60 and 130 ohms (inductive) on the long and medium wave-bands respectively.

(iii) Neon signs, traffic lights, etc.

The interference from neon signs may be either directly radiated or mains-borne. In these signs the current flows only during a short period of each cycle, as the ignition voltages and extinction voltages are high. In consequence the current wave is very peaky and considerable interference is generated. Some of the larger types of sign cover an extensive area and produce appreciable radiation. The disturbance caused by the radiated field is usually less than that due to mains-borne interference, and it is possible to adopt a remedy which is effective in both cases. Single-contact signs and multiple-contact signs produce interference of which the major portion is usually mains-borne. The interference from traffic lights is wholly mains-borne, as the metallic screening of all equipment in these appliances prevents radiation. The spectrum of interference from such plant is, as might be anticipated, similar to the general spectrum from spark discharge such as is shown in Fig. 9 referring to interference from overhead lines. As might be expected, the intensity tends to increase as the frequency of measurement is decreased, up to and beyond the limits of broadcast reception.

(iv) Converting and generating plant

The interference on the supply side due to motor-generators is that appropriate to large motors, which is normally small, although minor defects in commutation may cause abnormal interference. The same applies largely to the output side and to large rotary convertors and balancers, the interference from which is generally of the order of 1 or 2 millivolts except in abnormal instances.

Mercury-arc rectifiers form a recent development which has received special consideration, since they are frequently installed for local use during change-over from d.c. to a.c. supply as well as for distribution and traction. Normally either one pole is earthed or the rectifier feeds the outers of a 3-wire system, the neutral wire being earthed and supplied from a balancer. It is then found that a condenser connected between lines reduces the interfering voltage between either line and earth, and accordingly it is desirable to measure the asymmetric voltage between each pole and earth separately. Similarly, the impedance of the system supplied exercises an important influence. Measurements are therefore made either with the rectifier connected as in actual use, or with an h.f. impedance of 150 ohms, or the open-circuit h.f. voltage is measured. The latter is the highest voltage which normally occurs, and, if the internal impedance is known, the voltage across a known load may be computed. Fig. 27(*a*) shows the relation between the "maximum" voltage, the open-circuit voltage, and the short-circuit current, for one of the rectifiers tested. The "maximum" voltage is that obtained when the h.f. load resonates with the rectifier.

Table 8

INTERFERENCE VOLTAGES AT MAINS TERMINALS OF LIFTS

Type of lift	Level of loudest interference (db. above 1 μV) across 150 ohms			
	Symmetrical		Asymmetrical	
	900 kc.	180 kc.	900 kc.	180 kc.
(i) Hand controller, d.c. operation (driving motor at bottom of shaft)	90	90	90 98 (+) 88 (−)	58 58 62
(ii) Similar to (i), but driving motor at top of shaft	92	98	80	64
(iii) Similar to (ii)	88*	87†	94*	96†
(iv) Push-button type, d.c. operation (driving motor at top of shaft)	77*	84†	81*	85†
(v) Similar to (iv)	82	84	81	86
(vi) Push-button type, a.c./d.c. operation, motor-generator supply	40–50	42–56	40–50	42–56

* G.P.O. tests at 1 000 k.c. † G.P.O. tests at 190 kc.
Plus sign (+) indicates positive main to earth, minus sign (−) indicates negative main to earth.

Table 9

EFFECT OF GRID CONTROL ON INTERFERENCE FROM MERCURY-ARC RECTIFIERS

Degree of grid control*	0·1	0·5	0·7	0·8	0·9	1·0	Rectifier
Asymmetrical voltage (db. above 1 μV) at 200 kc.				51	50	37	Glass bulbs, 2/200 kW
	115	115	112	107	101	92	Glass bulbs, 20 kW
		92	92	92			Steel tank, 500 kW
Asymmetrical voltage (db. above 1 μV) at 1 000 kc.				45	40	15	Glass bulb, 100 kW
	100	97	94	92	90	87	Glass bulb, 20 kW
		75	71	72			Steel tank, 500 kW

* Ratio of output voltage with grid control to output voltage without grid control.

The interference is continuously distributed over the broadcast bands. With the steel-tank type, the h.f. voltage is usually only of importance in the long wave band, since the voltage decreases rapidly between 150 and 300 kc., being fairly constant for higher frequencies

[Fig. 27(*b*)]. The glass-bulb type may show an increase of interference with frequency at lower frequencies, but when it is connected to a load of low h.f. impedance, e.g. a cable distribution network, the higher frequencies are attenuated and the interference tends to follow a variation similar to that for steel-tank rectifiers. In such circumstances, the h.f. voltage across the feeders due to a non-grid rectifier does not exceed about 10 mV in the long wave band and about 1 mV in the medium wave band. Provided the rectifier is stable the magnitude of the load supplied has only a minor effect. The use of grid-control, however, causes an increase in interference as the degree of control is increased. Most of the increase occurs as the grid-control ratio varies from 1 to 0·8 (see Table 9); a higher degree of grid control does not give much further increase in interference.*

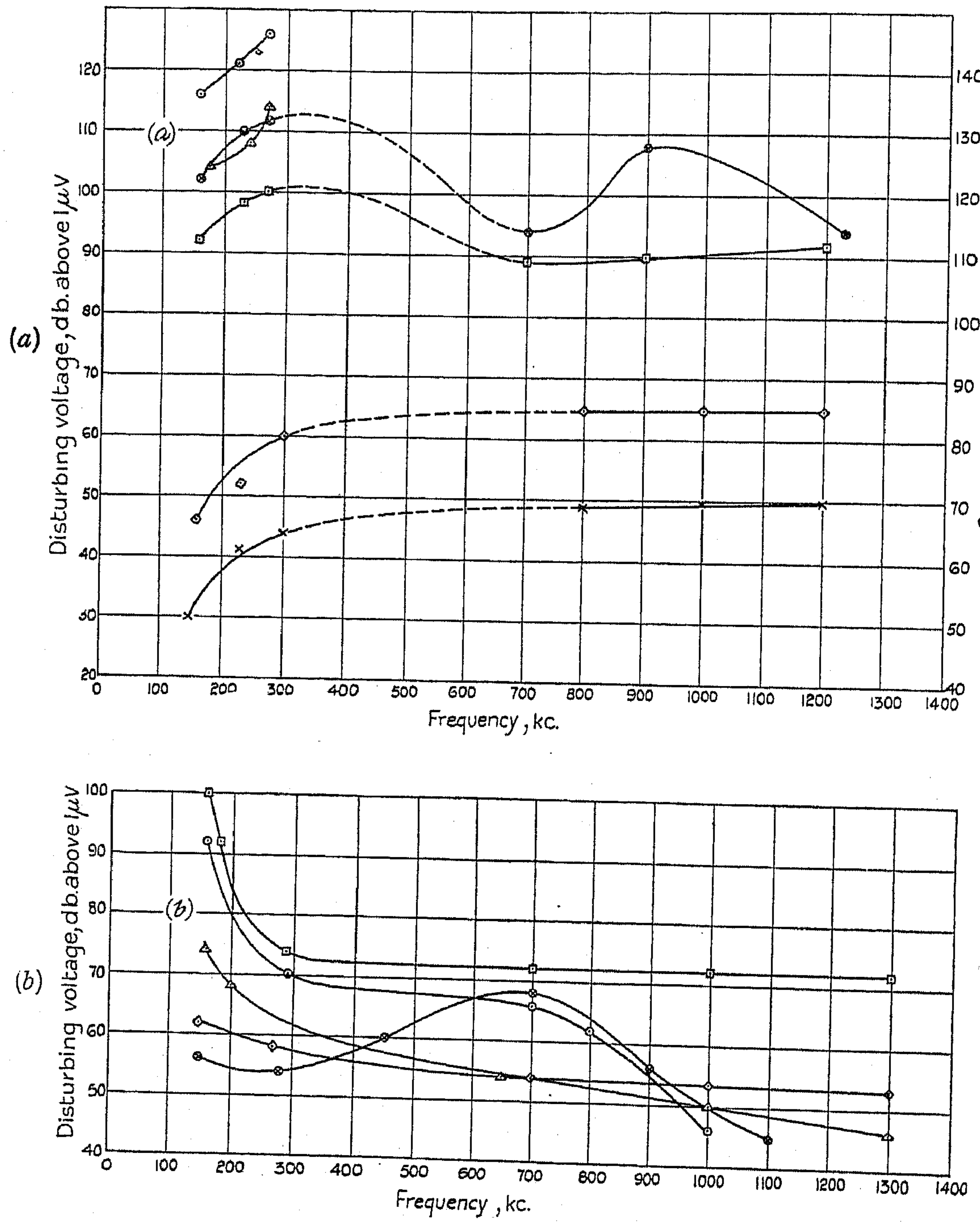

Fig. 27.—Interference from mercury-arc rectifiers.

(*a*) Maximum and open-circuit voltages and short-circuit current, for glass-bulb grid-controlled 6-phase 20-kW rectifier.

—— ⊕ —— ⊗ — Open-circuit voltage, grid-control ratio 0·7.
—— ⊡ —— ⊡ — Open-circuit voltage, grid-control ratio 0·97.
—— ⊙ —— ⊙ — Balanced voltage, grid-control ratio 0·7.
—— △ —— △ — Open-circuit voltage, grid-control ratio 0·97.
—— ◈ —— ◈ — Short-circuit current, grid-control ratio 0·7.

(*b*) Variation of interference with frequency.

—— ◈ —— ◈ — Asymmetrical (lines bonded), system load impedance, 4 glass bulbs (800 kW), grid-control ratio 0·89.
—— ⊗ —— ⊗ — Asymmetrical (lines bonded), load impedance 150 Ω, 2/3 phase, glass-bulb 5-kW rectifier without grid control.
—— ⊙ —— ⊙ — Positive line to earth, 150 Ω impedance, 300-kW steel-tank rectifier without grid control.
—— ⊡ —— ⊡ — Asymmetrical (lines bonded), load impedance 150 Ω, 500-kW steel-tank rectifier, grid-control ratio 0·73.
—— △ —— △ — Asymmetrical (lines bonded), load impedance 150 Ω, 4 glass bulbs (270 kW), without grid control.

The h.f. impedance of a glass-bulb rectifier is usually greater than 100 ohms and may be 400 ohms, whereas the impedance of the steel-tank type rarely exceeds 50 ohms. Accordingly, the reduction in interference due to a load of low h.f. impedance is greater in the former than in the latter case. This refers mainly to cable distribution networks where the h.f. impedance may be about 10 ohms. Where a small rectifier gives a local supply, the load impedance may be taken as of the order of 150 ohms, as for domestic items, and so the reduction in interference is not great.

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* The audio-frequency disturbance, i.e. the telephone harmonic factor, varies in a similar manner.

Small rectifiers may be supplied from the domestic a.c. mains, and then the interference on the a.c. side must be considered. It is usually much less (one-tenth to one-third) than on the d.c. side, and appears largely independent of the conditions on the d.c. side.

(3) METHODS OF SUPPRESSION

At the present time there is in this country no legislation protecting the licensed listener against radio interference, and a member of the general public who finds his reception interfered with or possibly spoilt has no redress apart, possibly, from applying in the High Court for an injunction for abatement of nuisance. The Postmaster-General, so far as his own commercial services are concerned, is in a somewhat better position, as he receives protection under the Telegraph and other Acts against interference with his services by power lines. As the commercial receiving stations are normally located well away from industrial or residential areas, many of the ordinary sources of interference are rarely troublesome and new overhead power lines are one of the few likely sources of interference.

In a number of other European countries legislation has been enacted with the object of protecting the broadcast listener from excessive and unnecessary interference with reception. This state of affairs is of considerable importance to British manufacturers of electrical appliances as, unless their goods are adequately suppressed as regards electrical interference, they may be refused entry into foreign countries. Moreover, foreign interference-free goods from countries with legislation may enter the British market and compete successfully with the unsuppressed goods of the British manufacturers. The question naturally arises whether it is essential, for the prevention of interference, to apply the remedy to the source of interference, or whether it is possible to modify the wireless receiver and aerial in such a manner that the interference is not received.

It is undoubtedly possible in many cases of interference to effect considerable relief by taking adequate measures at the receiving point. There remains, however, the larger proportion of cases where the interference is actually present as an electromagnetic wave of a frequency identical with that of the desired signal and for which no modification at the receiver can produce alleviation.

The Post Office has during recent years provided a free service to broadcast listeners for the purpose of dealing with cases of interference with broadcast reception. Where it is possible by modifications to the receiver or to the aerial system to avoid the interference, the listener is given advice accordingly. Where, however, no alleviation can be obtained in this way the source of interference is traced and, with the permission of the person owning the plant, the simplest and most economical method of suppressing the interference is ascertained. The owner is then persuaded, if possible, to adopt the remedy. It says a good deal for the public-spiritedness of electrical-plant owners in general that in the vast majority of cases very little persuasion is necessary to induce them to adopt suitable remedies.

A vast amount of work and annoyance would be saved, however, if some scheme of suppressing at least the smaller items of electrical plant at the manufacturers' works were adopted. The bulk suppression of interference in equipment, involves, however, a number of important considerations and requires some degree of international agreement, as otherwise plant which is suppressed sufficiently to meet the requirements of one country will not meet the requirements of some other country. It would be a great advance, therefore, if international agreement could be obtained as to a value admissible for the interfering noise voltage of an electrical machine or appliance when measured at the factory.

This value of admissible noise will depend on: (i) The h.f. field of the transmission to be protected. (ii) The degree of modulation of the transmission. (iii) The tolerable l.f. signal/noise ratio. (iv) The effective height of the listener's aerial. (v) The coupling between the receiving aerial and the interfering item. In regard to (i) and (ii), it is obvious that in the presence of a strong well-modulated signal from the desired transmission a given level of interference will produce far less annoyance than in the presence of a weak or poorly modulated signal. The transmission of music requires a wide range of modulation depths, which cannot, in general, be increased.

In connection with the work of the C.I.S.P.R. it has been agreed to consider the average field to be protected as being of the value of 1 mV per metre modulated at 80 %. Regarding the tolerable l.f. signal/noise ratio, it has already been mentioned in Section 1 (*a*) that as the result of tests by the C.I.S.P.R. the limiting value of this ratio has been found to be 40 db. when referred to the level of wanted signal at maximum modulation.

The fixing of an admissible value of noise is essentially a matter of compromise, as the choice of a very low figure would involve high cost in suppression services, whereas the adoption of too high a figure would relieve the cost of suppression at the expense of providing a higher signal field from the broadcasting service. In the latter connection there is a definite limit to the power of broadcasting stations, imposed by international agreement. The B.S.I. has agreed that, for frequencies between 200 and 1 500 kc., the maximum permissible interfering voltage at the mains terminals of a machine shall be 500 μV and that the interfering field strength shall not exceed 100 μV per metre at a distance of 10 yd. For example, if in a given case the effective height of aerial was 1 m. and the mains attenuation was 37 db., a field of 1 mV per metre would be adequately protected from mains-borne radiation.

(a) Suppression at the Source

The general principles of interference suppression are now generally realized, and in the case of the majority of smaller electrical appliances can be covered by general rules and specifications. The larger and more complicated electrical equipments, however, will generally require individual investigation and treatment for satisfactory suppression.

In suppressing interference at the source one of two courses may be pursued: either to prevent the generation of the interference, or, alternatively, if it is impossible to prevent the generation of the interference, to limit its effect by preventing its radiation or passage to the power supply mains. The first course can often be adopted in the case of interference from switching operations by preventing the sudden growth and sudden decay of current in the circuits in question. Circuits of the type used for spark-quenching are suitable for this purpose.

The majority of cases, however, do not lend themselves readily to this treatment, and it becomes necessary to prevent the interference leaving the source. This is done by screening or modifications of the unscreened portions of the circuit as far as radiation is concerned, and by the use of filter circuits to prevent the passage of the interference to the supply mains.

(i) Low-power items.

The methods to be adopted for the suppression of low-power equipment have been disclosed in earlier papers,* and the results of work and experience to date are embodied to a large extent in B.S.S. No. 613—1935.

A large percentage of cases can be cured by the use of condensers alone; for example, the items of equipment referred to in Table 7 were all suppressed down to a level of 200 μV of interference (46 db. relative to 1 μV), and this was accomplished by the use of condensers only in the percentages of cases given in Table 10. The use of condensers alone across the terminals of the appliance will be more efficacious where the internal impedance of the appliance to h.f. currents is high. The arrangement can be assumed to be equivalent to a potentiometer consisting of the machine impedance and condenser impedance shunted across the source of interfering voltage. The impedance of the condenser is low, so that the greater proportion of potential-drop occurs across the machine impedance and a very small proportion occurs across the condenser and supply mains. Care must of course be taken that the value of condenser chosen is such as not to resonate with the machine impedance at any frequency liable to cause interference, as in this event the interference may be accentuated. Furthermore, it is essential that the condensers with their associated connecting leads have as little inductance as possible, as otherwise their efficiency will be greatly reduced at the higher frequencies.

Table 10

Category	Percentage	No. of machines tested
A (*a*)	64	90
A (*b*)	73	34
A (*c*)	27	168
B	65	17

The I.E.E. Wiring Regulations recommend that all exposed metal on machines, when such metal is less than 8 ft. above the floor, shall be earthed. The Regulations also deprecate the use of 2-pin reversible plugs and recommend that for portable appliances the framework should either be of non-metallic material or protected with insulating material. As these Regulations are not always obeyed it is necessary to safeguard the users of such appliances from the possibility of electric shock.

When condenser suppressors are fitted to portable machines one condenser is usually necessary between the supply mains and one condenser between the neutral and the frame. With the use of 2-pin plugs it cannot be guaranteed that the latter condenser will remain connected as originally fitted, and it has therefore to be limited to a low value in order to avoid the risk of shock. This limitation has the effect of reducing the efficacy of the suppressors, but it has been found that if two 0·005-μF condensers are fitted in series across 250-volt a.c. supply mains and the centre point of the condensers connected to the framework, then a reasonable degree of suppression can be obtained for a large number of machines, and it has been shown by extensive tests that in the event of the user touching the framework and earth simultaneously no appreciable shock can be felt.

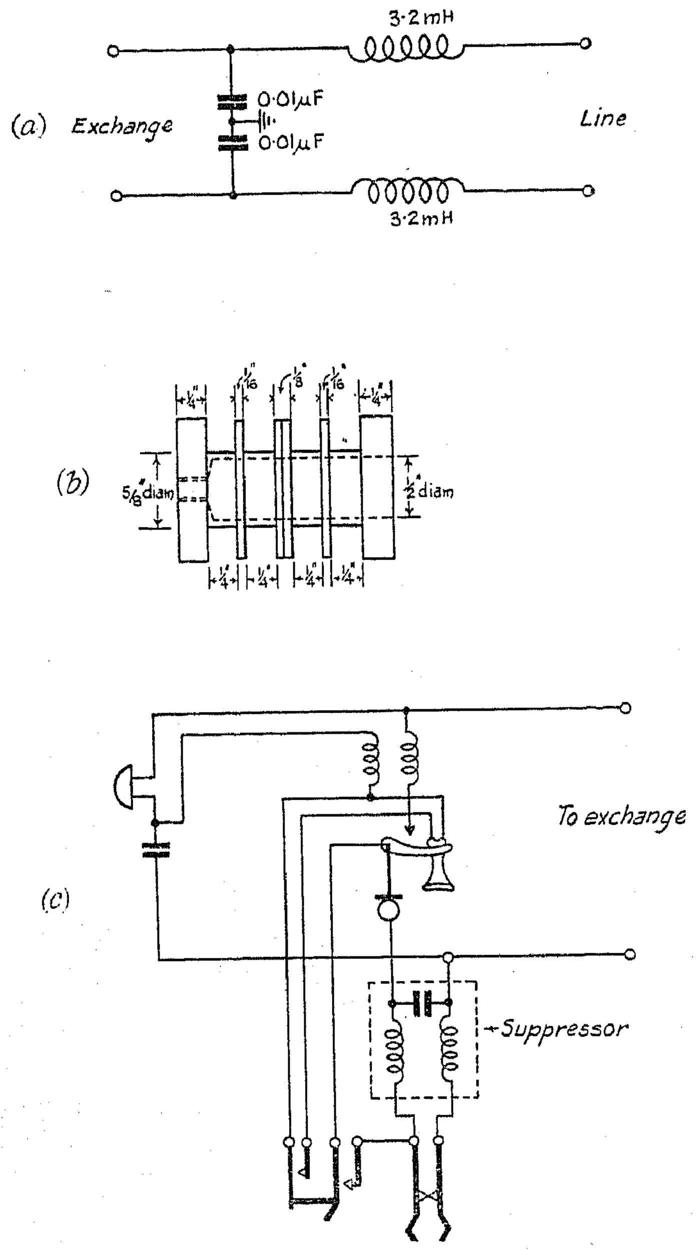

Fig. 28

(*a*) Line suppressor for automatic telephone exchange.
(*b*) Coil spool (moulded bakelite).
(*c*) Dial suppressor at subscriber's premises.

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* For example, A. Morris: *Journal I.E.E.*, 1934, vol. 74, p. 245; and *Proceedings of the Wireless Section*, 1934, vol. 9, p. 39.

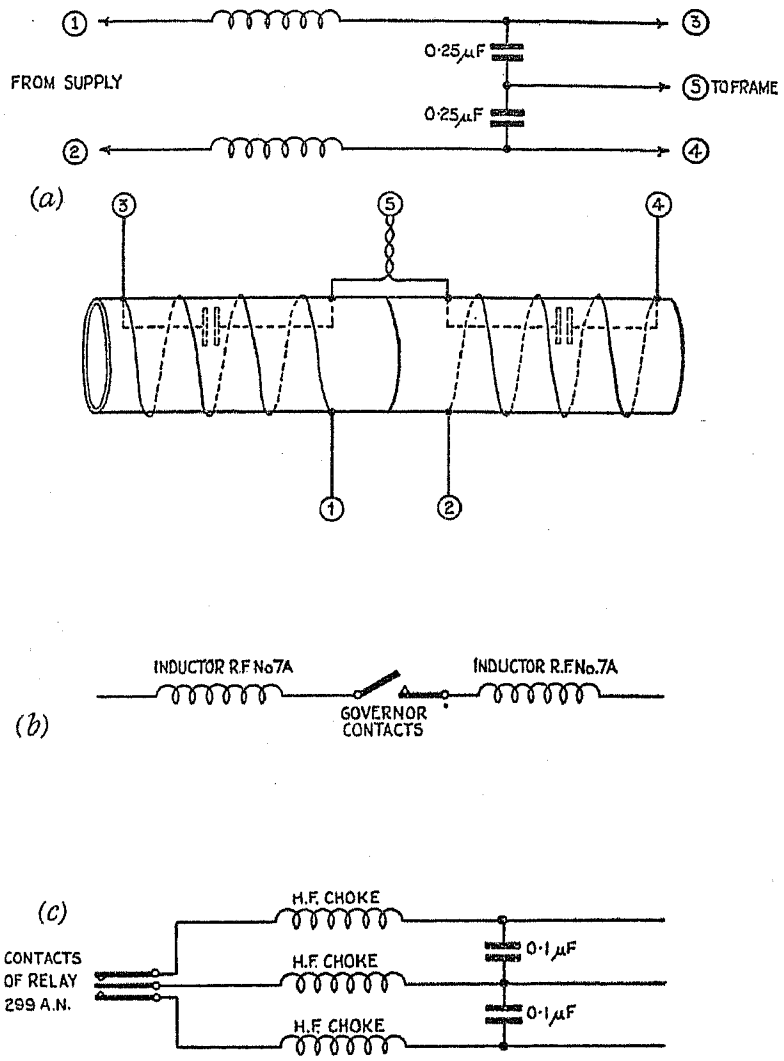

Fig. 29.—Interference suppressors for P.O. teleprinters.

(a) Mains input filter.
(b) Governor suppressor.
(c) Relay suppressor (for duplex voice-frequency circuit).

This method has been adopted as standard practice by the Post Office.

In the case of telecommunication apparatus it has already been mentioned that, as far as automatic telephone plant is concerned, the most effective method of suppressing interference from an exchange with overhead distribution is to provide filters on every circuit leaving the exchange building. Details of the line suppressors used for the purpose are given in Fig. 28(*a*). The impedance of the coils increased from about 3 500 ohms at 150 kc. to a peak of over 250 000 ohms at 1 400 kc., after which it decreased to about 100 000 ohms at 1 500 kc. The self-capacitance of each coil was about 5 $\mu\mu$F. Details of the coils used are given in Fig. 28(*b*). The suppressors are mounted in units for 50 lines and for 20 lines, and the overall dimensions of a 50-line unit are 1 ft. 7 in. high × 1 ft. 10 in. wide × 10 in. overall depth. The 20-line unit is similar, with an overall width of 1 ft. 3½ in. Particulars of the suppression achieved by this equipment are given in Table 6.

The interference sometimes experienced from dialling impulses can be alleviated by fitting a suppressor at the subscriber's dial. A typical arrangement of a suppressor for this purpose is shown in Fig. 28(*c*). The condenser is of 0·1 μF and is of the non-inductive tubular type. Coils of various types have been used, wound over the condenser. One arrangement consists of two coils of No. 36 S.W.G., d.s.c., having a loop inductance of 25 μH and a loop resistance of 0·88 ohm.

With regard to interference from machine telegraphs, this rarely causes trouble except in commercial receiving stations. One remedy in such cases is to install the machine telegraph apparatus in a screened enclosure

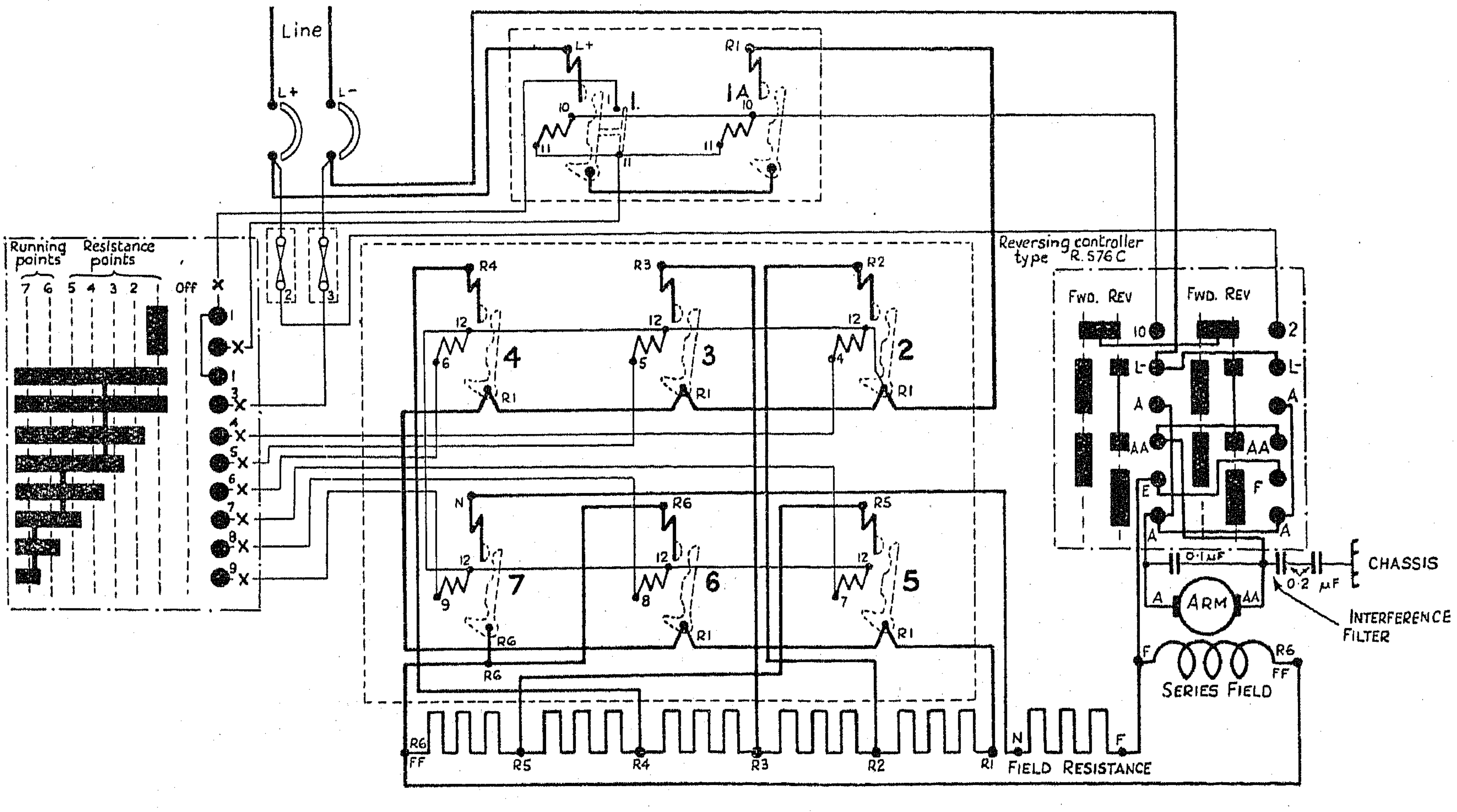

Fig. 30.—Schematic diagram of typical trolley-bus electrical equipment.

with filters in all incoming conductors. Another remedy is to provide suppression filters on the machine telegraph equipment. In severe cases it may be necessary to employ both remedies.

In the case of Post Office teleprinters installed at a radio receiving station, it was found necessary to fit suppressors to the mains supply, to the governor contacts, and to the sending and receiving relay. Details of the filters used are given in Fig. 29. The mains filter consists of two coils, each of 54 turns of No. 24 S.W.G. d.c.c. copper wire, pile-wound in six groups of nine turns on two tubular 0·25-μF non-inductive condensers. They can be fitted in the base of the teleprinter. The governor filter consists of two coils. A greater degree of suppression is obtained by screening these coils before fitting them within the base of the teleprinter. For the relay suppressor, commercial short-wave h.f. chokes are suitable as the current to be carried is small. This equipment was found to give suppression for the motor and governor interference over the band 150–19 000 kc.; less complete suppression might be sufficient in cases where the received signals are of the order of 100 μV per metre or more.

(ii) Trolley-buses.

The three main methods of suppression of the interference from trolley-buses are by treatment of individual items, suppressors on the overhead wires, and chokes in the main supply leads from the trolley arms.

In the first method, inductors are connected in each controller lead and condensers and resistors are connected across the controller contacts; there are also condenser filters on the driving motor, and on the brake compressor or exhauster motor, and inductors in the supply to the lighting motor-generator, if present. The details of the suppressors for one type of trolley-bus are shown in Fig. 30. The capacitance connected to the chassis does not exceed 0·1 μF, and it is considered that the shock obtainable from the discharge of this condenser would, under the worst condition, be inappreciable. Normally the leakage to earth from the chassis is adequate to prevent the potential of the latter from differing from that of the ground in the vicinity, but it is occasionally possible for the chassis to become effectively insulated from earth. By means of these filters, the interference from the items to which they are connected can be reduced to field strengths of the order of 20 db. above 1 μV per metre in the medium wave band, and 40 db. above 1 μV per metre in the long wave band, measured at a distance of 10 yd. from the vehicle and under the trolley wires. Individual suppression ratios as high as 80 db. have been reported in some tests. Table 11 shows some representative test-results for the suppression of interference due to controller operation.

Table 11

Frequency (kc.)	Type of vehicle	Nature of filters	Degree of suppression (db.)	At
300 and 1 100	Rheostatic braking	4-mH chokes	26	Town A
1 000 200	Company B	10-mH chokes on each side of each contact	53 30	Town B
900 200	Regenerative braking	ditto	43 33	Town C
900 200	Company C	ditto	54 21	Town D
900 200	Regenerative-rheostatic braking	ditto	28–42 29–33	Town E
900 200	ditto	ditto	36 36	Town F
900 200	Regenerative braking	5-mH chokes ditto	46 11	
900 200	ditto	As above, but with 0·1 μF across contacts	>72 54	
900 200	Regenerative-rheostatic braking	10-mH chokes on each side of each contact	40 30	Depot H
900 200		As above, but with resistances across certain contacts	80 60	Depot H
900 200	Mechanical braking only	10-mH chokes on each side of each contact	24 29	Town G
900 200	Air brake only	ditto	27 30	Town H
900 200	ditto Company D	ditto	43 42	Town J
900 200	ditto Company E	ditto	43 42	Town J
900 200	Eddy-current braking, Company F	ditto	42 40	Town J

In actual operation, interference due to collectors and due to the main contactors also occurs. The latter is usually small, as has already been noted, while the former is now much reduced by changes in design and maintenance. Accordingly the method will give adequate suppression both for radiation from the system and for direct radiation from the trolley-bus, except when the collector noise is unduly high, or when there are other sources of interference coupled to the overhead system, or in some of the exceptional conditions of weather, etc., already mentioned in Section (2).

If the lines themselves are loaded with condensers, the interfering currents pass to earth at adjacent condensers, while the impedance of the line is reduced, thus decreasing the voltage corresponding to a given interfering current. The first effect increases to some extent with the normal impedance of the line, which, however, can only vary within narrow limits unless special conductors are used, so that the effect will depend mainly on the resistance of the earth connection of the condensers. The second effect depends also on the internal impedance of the vehicle being greater than the external impedance. The efficiency of condensers thus varies according to the earth resistance of the standards to which the condensers are usually earthed and, to some extent, according to the vehicles in question. Tests on systems where the earth resistance was of the order of 10–20 ohms, as measured by ordinary earth-testers, indicated suppression ratios up to 40–50 db. in the long wave band and 20–30 db. in the medium wave band, with condensers of $0 \cdot 5\ \mu F$ connected to earth on each wire at 80-yd. intervals, the degree of suppression decreasing with increase of interval, and being about 20 db. in both wave bands for an interval of 350 yd. In some instances, adequate suppression can be obtained by condensers alone, as indicated in Table 12.

Table 12

SUPPRESSION OF TROLLEY-BUS INTERFERENCE BY LINE CONDENSERS ALONE AND IN CONJUNCTION WITH OTHER SUPPRESSORS

Trolley-bus suppression apparatus	Site	Distance of test position from overhead line (yd.)	Interference level (db. above 1μV per metre)	Condenser spacing (yd.)	Suppression (db.)	Frequency (kc.)	Remarks
Unsuppressed	A	10	47	440	20	1 000	Single trolley-bus, normal operation
Unsuppressed	A	10	42	80	25	1 000	Single trolley-bus, normal operation
Double-hump chokes ..	A	10	43	440	27	1 000	Single trolley-bus, normal operation
Unsuppressed			44	80		1 000	Single trolley-bus
Single-hump chokes ..	B	5	47	80		1 000	Normal service
Single-hump chokes ..	B	5	29	80		1 000	Single trolley-bus
Unsuppressed	A	10	32	240	30	200	Single trolley-bus
Unsuppressed	A	10	19	80	43	200	
Single-hump chokes ..	A		32	240			Normal service
Double-hump chokes ..	A	10	34	440	28	200	Single trolley-bus
Single-hump chokes ..	B	5	44	80		200	Normal service
Unsuppressed	B		34	80			Single trolley-bus
Single-hump chokes ..	B	5	33	80			Single trolley-bus

Suppression by condensers alone may not always be adequate in severe instances, but condensers may conveniently be used in conjunction with other methods since they eliminate collector noise and line noise from other sources, and the failure of a unit does not invalidate the suppression since the influence of the remaining units is exercised over the whole of the line. In this way the frequency of severe "clicks" may be considerably reduced, to the point at which they become tolerable, while systems or sections which are the source of strong interference may be isolated from the systems and sections serving residential areas where the streets are narrow.

If choke coils are connected in the main supply leads to the trolley-bus equipment, the interfering voltages generated therein are prevented from reaching the line while currents due to collector disturbances are prevented from circulating in the system. These chokes must carry a heavy current, however, so that their design forms a difficulty on account of the non-paying load and the reduction in maximum tilting-angle which may be caused. The latter objection may be minimized by winding the coils with aluminium strip in lieu of copper. They are usually mounted on the roof, but may be mounted in the cab without very much loss in efficiency. Existing designs are mainly of two types, the single-hump type, giving a maximum impedance in one wave band, and the double-hump type, having an auxiliary circuit (Fig. 31) such as to produce two maxima of impedance, one in each wave band. Suppressions of 20–30 db. have been obtained by these means, but displacement of the maxima from the test frequency and severe collector noise may considerably reduce the suppression.

The present conclusion is that for severe cases, where the streets are narrow, a combination of suppressors on each disturbing item on the trolley-bus, with a few line condensers at special situations such as turning circles, crossings, and section points, is desirable. If the line is in bad condition or if wheel collectors are used, more line condensers may be necessary. In other cases the line condensers may be omitted, or line condensers may often be employed alone, if at sufficiently frequent intervals. For wide streets where aerials can be located at, say, 50 yd. from the route, main choke coils may be used. Where these are already in use the suppression they give may, if inadequate, be suitably reinforced by the addition of line condensers.

(iii) Electric lifts.

The component items and circuits of electric lifts are treated separately in this paper, and the methods applicable have, in many instances, already been described; such as, for example, the main motor, the motor-generator, convertor, or rectifier in a.c. and a.c./d.c. systems; the gate-closing motors, bells, indicator lights, fans, and similar auxiliary equipment. The control circuits show some special features. If these circuits are operated from a panel in the car the interference must be prevented from flowing in the trailing cable and being radiated therefrom. To this end, chokes of about 6 mH should be connected in each lead, both at the panel in the car and at the controller where the relay coils are situated. In addition, condensers of 0·5 or 1 μF may be connected from each lead to earth on the trailing-cable side of the chokes. A less but often adequate degree of suppression can be obtained by mounting the chokes at the halfway box instead of in the car itself, while the filters at the controller panel or, at least, the chokes therein comprised, may often be omitted. In general, the more rigorous treatment is required in control circuits energized from the full mains voltage since these circuits are more highly inductive. In many modern lifts a.c. controls are used and need little suppression.

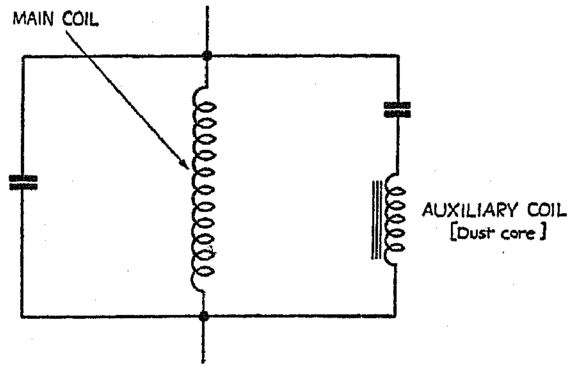

Fig. 31.—Schematic diagram of trolley-bus double-hump choke.

Control circuits, including gate interlocks, which do not enter the trailing cable are treated in a similar manner, filters being required at the operating switch and at the relay coil. If the wiring is in conduit or screened the chokes may often be omitted from the filters. Screened trailing cables have not yet come into general use.

The tripping, holding, and closing coils of the circuit-breakers and contactors should be furnished with condensers of 1 or 2 μF, connected between each end of the coil and earth. In some older types of lift these coils are directly controlled from the car, the circuit being included in the trailing cable. Severe interference is experienced in such circumstances, and the methods described previously for control circuits must be applied. It is also often advantageous to connect condensers up to 4 μF across the switch contacts in the car. The brake-magnet coil, which may also be a source of interference, can be suppressed to the extent of about 40 db. by means of an arc-suppressor of the rectifier type, as is often employed in inductive circuits. Such devices form an alternative in many instances to the condenser and choke-condenser filters described. The metalwork of the shaft should be well bonded to earth, since appreciable voltages may arise along the length of the shaft.

The interference passing into the supply mains, if the domestic supply or a system coupled thereto is employed, can be satisfactorily reduced by a mains filter of the normal choke-condenser type. Table 13 shows the methods and extent of suppression which have been achieved in some typical instances.

(iv) Ignition systems.

As regards the normal broadcast wavelengths, condenser suppressors of a conventional nature are applied to the various items of electrical equipment of automobiles, such as the dynamo, coil or magneto, windscreen wiper, horn, etc. In addition, screened cable may be used for the wiring, the screen being bonded to the chassis. Such measures are mainly of importance for car radio where, in addition, or alternatively, the aerial and lead are screened if in proximity to the car wiring.

The spark system itself affects receivers along the route or in the neighbourhood, and requires special consideration. Complete screening is entirely effective but, at present, expensive and mainly employed in aircraft and for military and police purposes. The coil or magneto is encased in a low-resistance box to which is attached, by a metallic gland, the screened cable leading to the distributor, similarly encased in a metal box. The plug leads are also formed of screened cable, attached to the distributor and the plug by metallic glands completely enclosing the plug head and distributor terminals. The screened cable is bonded at intervals to the chassis.*

Automobiles may be partially screened by using screened and bonded ignition leads or by using a metal cover substantially enclosing the high-voltage circuit. It is then necessary to apply a choke-condenser filter at the battery terminals of the coil. This scheme appears sometimes to give adequate suppression both for car-radio and for external receivers. One serious disadvantage with screening is the capacitive load on coil or magneto, which may reduce its voltage unless it is specially designed, and decrease the life of the contact point.

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* See Air Ministry Specification DTD GE 125, Issue 6, and B.S. Specification in course of issue.

Partial screening may in many cases meet this objection,* but difficulties will occur if existing coils are used with long-gap sparking plugs. It seems possible, however, that with these plugs certain other alleviating factors may play a part.

Table 13

DEGREE OF SUPPRESSION OF INTERFERENCE FROM ELECTRIC LIFTS

(*a*) *Radiated*

Type of lift (see Table 8)	Suppression condition	Level after suppression (db. above 1 μV per metre) at 10 ft. from C/L of lift	
		900 kc.	180 kc.
(i)	Chokes at panel	70	46
	Chokes at halfway box	62	47
	Chokes plus condensers (earthed) at panel	62	42
	Chokes plus condensers at car controller	48	40
	Do., plus chokes and condensers at panel	48	40
(iii)	Chokes at panel plus choke/condenser filter in mains	33	34
(v)	Chokes plus condensers (earthed) at halfway box	40	23
	Same filters at controller panel	36	21
(vi)	Chokes plus condensers (earthed) at panel (gate motor leads) plus chokes and condensers at motor switch contacts ..	16†	25‡

(*b*) *Mains Interference*

Type of lift (see Table 8)	Suppression conditions in supply leads	Level after suppression (db. above 1 μV)			
		Symmetrical		Asymmetrical	
		900 kc.	180 kc.	900 kc.	180 kc.
(i)	Chokes with condensers (earthed) on line side	52	62	45	46
(iii)	Chokes at panel plus choke/condenser filter in mains	39†	26‡	44†	28‡
(v)	Chokes with condensers (earthed) on line side	38	36	45	44
(vi)	None	45	48	45	48

† G.P.O. tests at 1 000 kc. ‡ G.P.O. tests at 190 kc.

On account of the difficulties mentioned, alternative methods have been studied which fall into two types—resistors, including resistor-condenser combinations, and choke-condenser units. Resistors must be placed as close as possible to the source of the spark, i.e. at the actual plug and distributor terminals. With a single-cylinder magneto or with a brush-type distributor, the resistors need only be mounted at the plugs. With magnetos, the distributor is usually incorporated in or close to the magneto so that no resistor is necessary in the distributor lead, but such a resistor is necessary when a long lead connects a coil to a distributor. The use of long distributed resistors as the plug leads themselves avoids duplication of resistors at each end and is more efficient for the same resistance value. In addition, the efficiency can be very considerably increased by covering a short length of the lead at the plug end by a screen of metal foil, as in Fig. 32, so as to form a small condenser of about 20 $\mu\mu$F, the foil being bonded to the cylinder block. This forms a network giving a high attenuation from the plug for very high frequencies. Table 14 indicates the degree of suppression which may be attained. The suppression necessary depends, with ultra-short waves, on the service field-strength envisaged and the location of the receiver. In general, it appears from Section (2) that suppressions of not more than about 25 db. should prove satisfactory in the great majority of cases.

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* With some car designs it is possible to enclose the whole of the ignition circuit in a single screen bolted to the engine.

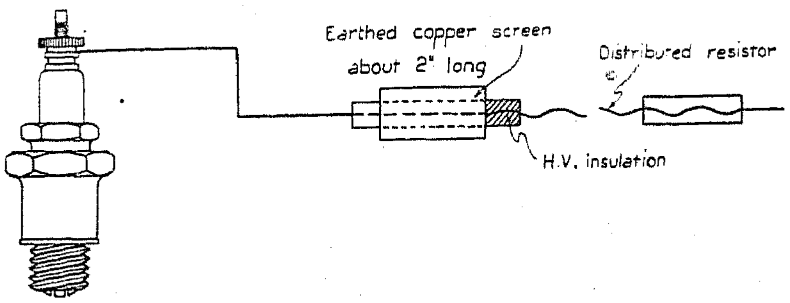

Fig. 32.—Distributed resistor, with screen, for ignition systems.

The figures in Table 14 were obtained partly on a model ignition system and partly from tests on a number of automobiles of various ages and sizes produced by the principal motor-car manufacturers. Often the degree of suppression required, after the ignition layout has been suitably arranged, is such that only a resistor in the distributor lead is necessary.

Table 14

Frequency (Mc.)	Resistance (10³ ohms)	Suppression (db.)
Single-cylinder magneto		
12	100†	27
	50†	27
	25†	24
	25‡	34
45	25†	17
	25‡	23·5
	25*	16
Coil and 4-cylinder brush-type distributor		
21	25†	22
45	25†	25

Frequency (Mc.)	Resistance (10³ ohms)		Suppression (db.)
Coil and spark-type distributor			
	Distributor lead	Plug leads	
12	20*	25†	29
	—	50†	30
45	20*	25†	20
	20*	30‡	40
	15*	10*	25
	5*	5*	15

* Carbon resistor.
† Distributed resistor.
‡ Distributed resistor with screen.

If h.f. chokes are employed at the plugs, it is essential to incorporate the small condenser, already mentioned, on the lead at its point of attachment to the choke, as is the case with the chief proprietary form of this suppressor. The suppressions thereby obtained by the use of suitable components have varied from about 18·5 db. at 50 Mc. to 30 db. at 12 Mc.

(v) Mercury-arc rectifiers.

With some exceptions, satisfactory suppression of the interference on the d.c. side is obtained by means of condensers connected between each output lead and earth or between the anodes and cathode and earth, or by a combination of these methods. By means of such condensers, of capacitance up to 1 or 2 μF, the disturbance can usually be reduced to 600 μV or less. Since a reduction of 10 db. may be allowed for attenuation in the cable to the nearest consumer, this corresponds to about 200 μV applied in a manner comparable to that associated with a domestic item.

With other types of system, with certain rectifiers of the grid-control type and with rectifiers not completely balanced or stabilized, a higher degree of suppression is desirable. This may be achieved by h.f. chokes inserted in each feeder, with condensers connected between the load side of each choke and earth. The inductance of these chokes is limited by the load current to be carried. Standard values for air-core chokes are 150 μH, 300 μH, 800 μH to 1 mH. The use of a choke with an iron core in shell form is of advantage as regards size and carrying capacity when dealing with heavy currents. Table 15 shows the methods of suppression adopted in a number of particular instances. A wide variation of efficiency of suppression is observed, and shows that regard must be paid to the type of load and relative impedances of load and rectifier in choosing the best methods of suppression. Item (2) in Table 15 is an example of this. The low suppression-ratios achieved with chokes in this instance are due to the fact that, owing to the low h.f. impedance of the system supplied, the insertion of chokes increased the interfering voltage at the terminals of the rectifier so as largely to neutralize the attenuation provided by the choke-condenser filter applied.

In conditions obtaining in this country a disturbance on the a.c. side needs treatment only in those cases where the domestic mains system is used and where the interference level of the rectifier is unusually high. Suppression can then be achieved by a filter of normal type with a choke in each phase and a condenser of 0·1 to 1 μF connected between the supply side of each choke and earth.

(vi) High-frequency medical appliances, and cases where screening is essential.

Although the suppression of the high-frequency interference arising from the operation of diathermy and similar types of apparatus is quite simple, the practical application may at times present some difficulty where apparatus is already permanently installed. It is therefore important that in all new hospitals and medical institutions provision should be made during construction for prevention of radiation by the means described later.

Two methods are available. The first is to redesign the apparatus in such a way that the secondary forms an output circuit which in linear dimensions is as small as possible, and which has a metallic return to reduce earth currents to a minimum. This method has so far not been adopted to any large extent, principally on the grounds of expense, but an experimental violet-ray set has been designed by the Post Office to these requirements and is practically non-interfering. The second method, which can be adopted for complete suppression, is the enclosure of the apparatus and patient in an earthed metallic screen and the provision of h.f. filters

Table 15

SUPPRESSION OF DISTURBANCE FROM MERCURY-ARC RECTIFIERS

Type of rectifier	Capacity	Type of load	Condition of test	Suppression device	Frequency (kc.)	Suppressed level (db. above 1 µV)	Suppression ratio (db.)
(1) 6-phase, steel tank, non-grid	500 kW 460 V	Cable distribution system	Positive line to earth	2 µF each anode and cathode to earth, and on balancer	217	48	33
					900	39	26
(2) 6-phase, steel tank, non-grid	300 kW 480 V	Cable distribution system	Positive line to earth	12 µF each line to earth	150	52	— 1
					1 000	42	3·5
			Negative line to earth	12 µF each line to earth	180	49	8
					1 000	46	1
			Asymmetrical (lines bonded) to earth	1 µF each anode and cathode to earth, and between lines	180	55	9
					1 000	48	0
			Positive line to earth	150 µH in positive line, 12 µF to earth	180	49	2
					1 000	41·5	4
(3) 6-phase, steel tank, grid control	500 kW 480 V	Cable distribution system	Line to earth	0·1 µF each anode and cathode to earth, and between lines	180	82·5	12·5
					1 000	62·5	11·5
			Line to earth	200 µH in each line, 10 µF to earth on load side	180	47	25
					1 000	56	10·5
(4) Glass bulb, non-grid, two 3-phase units	480 V 10 A	Local supply	Line to earth	1 µF each anode and cathode to earth, and between lines	205	84	— 10
					900	64	6
			Line to earth	1 µF each line and neutral to earth	205	19	54
					900	30	40
(5) Glass bulb, non-grid (2 units, giving 12 phases)	270 kW 480 V	Cable distribution system	Line to earth	1 µF each line to earth	220	40·5	28·5
					1 000	36	11
				1 µF anodes and cathode, and between lines	220	27·5	40·5
					1 000	—	Order of 50
(6) 6-phase, glass bulb, grid control (0·89), 25 cycles per sec.	2 units gave 230 V, 200 kW	Cable distribution system	Line to earth	1 µF each line to earth	180	40	15
					1 000	49	6
(7) 6-phase, glass bulb, grid control (0·97), grid control (0·1)	360 V 60 A	Local supply	Line to earth	400-µH chokes in each line, 0·5 µF each line to earth	217	51	46
					900	58	32
				400-µH chokes in each line, 0·5 µF each line to earth	217	59	51
					900	58	42

Table 16

System	Cost per sq. ft.	Screening for walls, door, ceiling, and floor	Wire-mesh reinforced glass for window 3 ft. × 4ft.	Mains filter	Labour costs	Total cost
Metallized paper	½d.	£1 10s.	12s.	£4	£5	£11 2s.
Zinc-sprayed paper ..	2d.	£6 0s.	12s.	£4	£5	£16 12s.
Zinc-spraying on wall ..	1s., including labour	£31 10s. Floor covered with wire netting	12s.	£4	£1	£37 2s.
¼-in. galvanized netting ..	1d.	£3 10s.	12s.	£4	£5	£13 2s.

on all circuits entering the screened enclosure. It is also necessary that all water pipes, gas pipes, conduits, etc., be efficiently bonded to the screen.

Instances may occur in which the nearest broadcast listener, although situated outside the field of directly-radiated disturbance, is yet within the area over which receivers may be affected by the h.f. components propagated via the supply mains. Provided that the electric supply system in the locality is laid underground an adequate degree of suppression should be gained by the insertion of an h.f. filter at the point of entry of the supply mains to the premises in which the interfering apparatus is housed.

In the case of a violet-ray set used in a private house or in a hairdresser's establishment, a screening cubicle may be constructed very simply by stretching ¼-in. mesh galvanized-iron wire netting over a wooden frame and bonding this to similar netting laid on the floor. A cubicle having the dimensions 7 ft. × 6 ft. × 6 ft. 6 in. would be of suitable size and could be erected and draped to taste at moderate cost. In the case of a hospital or medical practitioner, however, it is usually more convenient to screen the room in which the electrical treatment is given. This may be effected by the use of any one or a combination of the following screening materials: metallized paper, zinc-sprayed paper, zinc spraying on the walls and ceiling, or small-mesh galvanized-iron wire netting. If the room to be screened is in course of construction it may be found practicable to employ wire mesh for screening purposes, as this can be embedded during the process of plastering, and when finished the screening would be hidden.

Metallized paper is initially very effective, but in one case its efficiency was found to decrease considerably with age, the resistance between adjoining strips of paper at pasted joints increasing as the paste dried out. It follows that if metallized or zinc-sprayed paper is used, special attention must be given to the permanency of the joints. Owing to the high cost of screening by spraying zinc directly on to the walls, this method has not been tried out in practice on an actual room. From tests carried out on small specimens, however, it would appear to be satisfactory.

As a guide to the relative costs of the various methods, the estimated costs of screening a room 12 ft. × 14 ft. × 8 ft. high are given in Table 16. The cost of a 15-ampere h.f. mains filter is also included in each case.

For decorative purposes, paint or wallpaper could be used to cover the screening in the case of the first three methods referred to in Table 16. The cost of this is not included. In the fourth method the wire netting can with advantage be embedded in the plaster of the walls and ceiling. If it is embedded during construction of a new building there is no additional cost. If replastering is necessary this will involve an additional cost of about £10. The metallized paper consists of

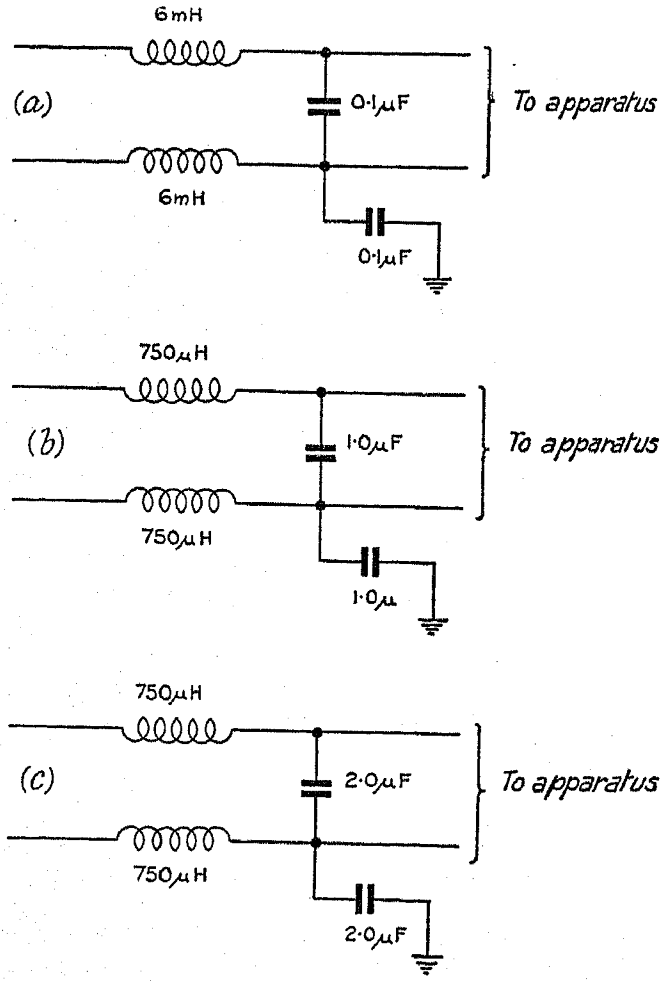

Fig. 33.—Suppressors used in tests on electromedical apparatus.

Table 17

PORTABLE HIGH-FREQUENCY MEDICAL SETS

Type of set	Frequency of measurement	Mains interference voltage measurement*				Radiated field strength				
		Suppressed (S) or unsuppressed (U)	Voltage values (db. above 1 μV)		Suppressors, Type (Fig. 30)	Type of screen†	Values in db. above 1 μV per metre			
			Symmetrical	Asymmetrical			3 yd.	10 yd.	20 yd.	30 yd.
	kc.									
Portable violet-ray	190	U	74	102	(*a*)	None	81	40		
		S	39	37		C	42	5		
	800	U	85	87		None	74	38		
		S	36	38		C	34	2		
	1 200	U	87	88		None	59	36		
		S	40	38		C	17	0		
Portable violet-ray	190	U	68	74	(*a*)	None	79	49		
		S	31	33		C	33	0		
	800	U	65	70		None	60	37		
		S	34	34		C	26	0		
	1 200	U	78	86		None	53	31		
		S	38	42		C	20	6		
Spark diathermy, ultra-short wave, output 2·5 amp.	190	U	73	76	(*b*)	None		44	40	37
		S	33	37		C		18	16	12
						B		30	10	—
	800	U	72	78		None		50	45	39
		S	33	35		C		28	24	18
						B		35	26	—
	1 200	U	101	107		None		65	58	47
		S	34	42		C		40	36	24
						B		46	36	—
Spark diathermy, medium wave, output 2·8 amp.	190	U	89	89	(*c*)	None		50	42	24
		S	33	33		C		29	23	<20
						B		35	24	22
	800	U	82	88		None		56	50	34
		S	42	46		C		36	31	<20
						B		39	29	27
	1 200	U	77	84		None		59	52	45
			37	27		C		40	34	26
						B		42	32	28

* In the mains-voltage measurements the frames of the portable machines were not earthed, but the frames of the diathermy machines were earthed.

† Type of screen B: Galvanized-wire mesh applied to a room 16 ft. × 10 ft. × 8 ft.
Type of screen C: Paper-backed aluminium foil, metal 0·010 mm. thick, applied to a room 16 ft. × 10 ft. × 8 ft.

aluminium foil, 0·005 mm. thick, with a paper backing, and is applied in the same way as wallpaper with the foil outwards. There should be a "lap" of at least 1 in. at joints, and to ensure continuity a 2-in. strip of aluminium foil is pasted over the joints.

The interference from a number of types of portable electrotherapy or "violet-ray" apparatus and diathermy apparatus on the long-wave and medium-wave broadcast bands has been investigated. The levels of interference on the supply mains were measured with and without suppressors, and under the symmetrical and asymmetrical conditions. The radiated field-strength was also measured at various distances under the screened and unscreened conditions. Typical measurements are given in Table 17. The types of h.f. filters used in the tests are shown in Fig. 33.

(b) Measures at Listener's Premises

(i) Type of aerial.

Wherever possible an outdoor aerial should be used, as this not only possesses a lower coupling with the supply mains but at the same time can have a much greater effective height than the indoor aerial. Frame aerials of the small type associated directly with the receiver are inefficient compared with the outside aerial, and although they possess certain directional properties these are rarely effective against interference because the frame aerial when rotatable on a vertical axis is directional only to vertically-polarized waves, and while the interference waves are largely vertically polarized there will be normally sufficient interference polarized in other directions to render the frame aerial of little avail. The most satisfactory arrangement will comprise an aerial well removed from the source of interference and connected to the receiver by some type of feeder system.

The feeder may be of the concentric or balanced 2-wire type. Where the feeder has to pass through an appreciable interference field as, for example, in going from top to bottom of a large building such as a block of flats, the single-core type of cable may not give such effective protection as a balanced type of cable, although much depends upon the actual circuits used in each case and the type of interference encountered. Enclosing a conductor in a metallic shield will not prevent its being affected by travelling waves of radiated energy; the shield will, however, protect the inner conductor from capacitive coupling to the electric supply mains, which is often a serious cause of interference. In the balanced transmission line both conductors are equally affected by radiated energy, but suitable terminal arrangements enable these effects to be balanced out. An arrangement of this kind is shown in Fig. 34(*a*). Here the aerial circuit includes the primary of the aerial transformer, the secondary of which feeds the balanced line. If the line is subject to radiated interference this will produce identical e.m.f.'s in each line conductor, which will set up currents between the mid-points of the line windings of the transformers and earth, but these currents will not introduce energy into the aerial or into the receiver.

A modification of the circuit shown in Fig. 34(*a*) is sometimes adopted, to allow the feeders to be used as the earth connection of the aerial itself. This arrangement is shown in Fig. 34(*b*), where the earth terminal of the primary of the aerial transformer is connected to the mid-point of the secondary. This circuit has the defect that any interference induced in the feeders in parallel enters the aerial through the primary of the aerial transformer, thereby producing loop currents which affect the receiver. This scheme, therefore, is not to be recommended in all cases, and it is preferable that the aerial, if possible, be earthed well away from the field of interference, as in (*a*). Where, however, the source of interference is capacitive coupling to the feeder the system will be effective if an earthed metallic screen is provided around the cable. Where the aerial can be placed so that the feeder can be buried or earthed along its route, the single-core concentric type will be suitable. Fig. 34(*c*) shows an arrangement of this kind.

One of the difficulties in connection with the provision of remote aerials and feeder systems arises from the increasing use of the so-called all-wave receivers which operate over a range of 150–30 000 kc. It is not easy to produce a feeder system which will operate satisfactorily over this wide range.

If the interference is most serious on the short waves an arrangement can be adopted which works as a balanced system on short waves and as an unbalanced system on medium and long waves. An arrangement of this kind is shown in Fig. 34(*d*), where the aerial consists of a dipole with a coupling coil in the centre to which the feeders are tapped: L_1 and L_2 are respectively the primary and secondary windings of a short-wave transformer, C_1 and C_2 are condensers of values of the order of 50 $\mu\mu$F, while L_3 is an inductance of about 50 μH. At very high frequencies the aerial behaves as a dipole feeding the balanced line. At the receiver end of the line the loop currents set up a voltage across the secondary of the transformer, one end of which is connected to the

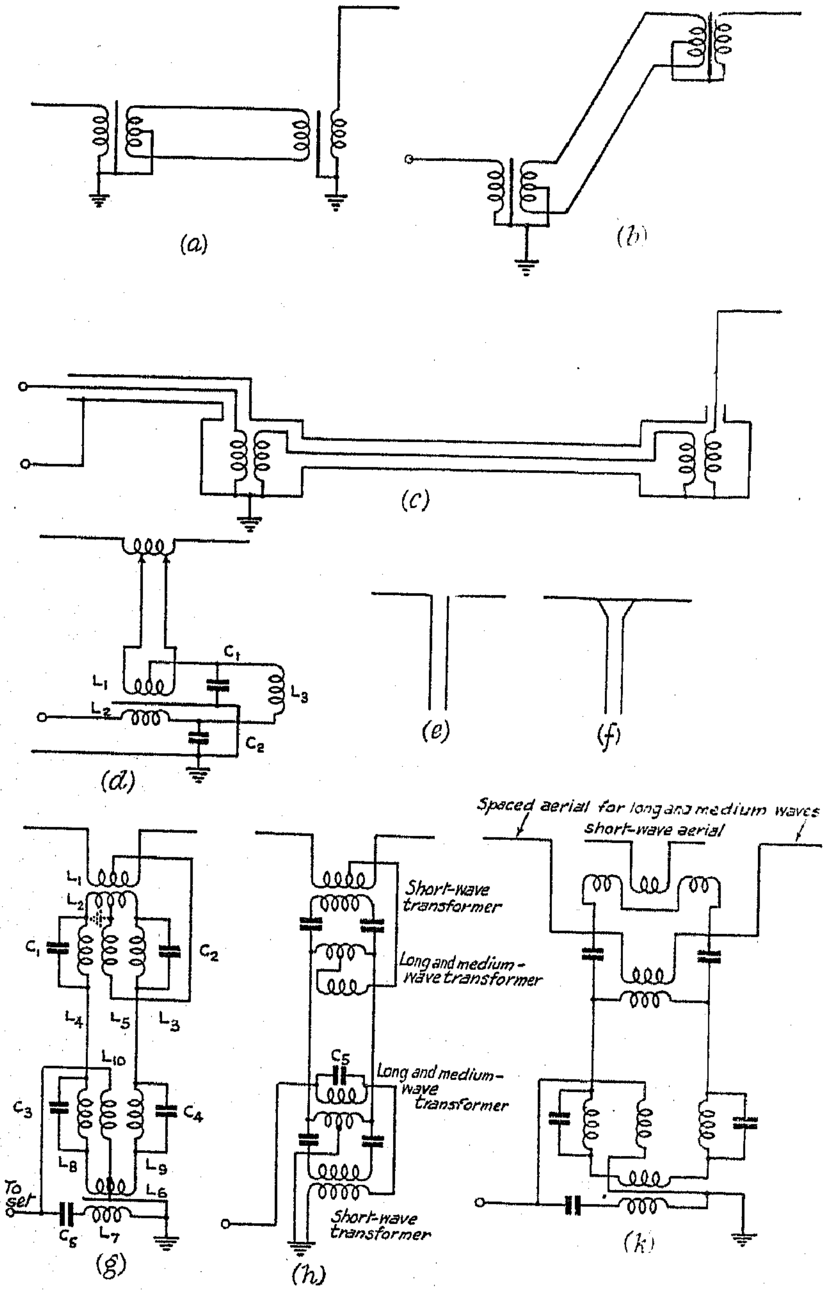

Fig. 34.—Aerial and feeder arrangements.

receiver while the other is effectively earthed through the condenser C_2. Short-wave interference induced in the line produces currents in the same direction in each of the feeders. These currents pass to earth through C_1. At these frequencies L_3 has a high impedance, so that very little of this interference passes to the set. At medium and low frequencies the dipole picks up very little energy as a balanced aerial, and a larger proportion of energy can be obtained from the dipole and feeder acting as a T aerial. This energy produces no output from the transformer, but as C_1 and C_2 have high impedances at medium and low frequencies the energy passes to the set through L_3, which has a comparatively low impedance. Alternative methods of connecting the dipole to the feeders are indicated in Figs. 34(*e*) and 34(*f*). This arrangement will not prevent the entry to the receiver of interference picked up by the antenna and feeder on medium and long waves.

Where an all-wave feeder system is of appreciable length it may be desirable to terminate it with transformers for long and medium waves as well as for short waves. An arrangement for this purpose is shown in Fig. 34(*g*). Here L_1 is the primary of the short-wave aerial transformer connected to a dipole aerial as shown in Fig. 34(*d*). The secondary winding is L_2, and both windings are centre-tapped. Across the centre taps is connected L_5, the primary of the long- and medium-wave transformer, having balanced secondaries L_3 and L_4 shunted by condensers C_1 and C_2 to allow passage for the short-wave line currents. The connections at the set transformer are somewhat similar. Here L_{10}, the secondary of the long- and medium-wave transformer, acts as a choke across the set on short waves, while C_5 presents a high impedance to long and medium waves. Other arrangements are possible; for example, the line windings of the transformers may be arranged in parallel instead of in series at one or at both ends of the line, while the secondaries of the set transformers may be connected in series instead of in parallel. One such possible arrangement is shown in Fig. 34(*h*).

The condensers C_1, C_2, C_3, and C_4 in Fig. 34(*g*), and C_5 in Fig. 34(*h*), can be arranged to be provided by the self-capacitances of the long- and medium-wave windings, but it is probably better that these windings have a low self-capacitance and the additional capacitance required be provided by small adjustable condensers. The condensers can then be adjusted to cover any capacitance unbalance in the windings.

None of these systems protect the feeders from radiated interference on long and medium waves, for the reasons previously mentioned. Such protection can be afforded by using a separate earth for the primary windings of the long- and medium-wave aerial transformer, as indicated in Fig. 34(*a*). The dotted connection, Fig. 34(*g*), indicates how this can be done. It is preferable to isolate completely the secondary circuits of the aerial transformers from the primary. The efficacy of the arrangement will of course depend on the removal of the earth connection from the field of interference.

If this cannot be arranged some form of balanced aerial for these wavelengths must be adopted. To be entirely effective, the aerial should consist of a loop or of two symmetrical spaced aerials, as, for example, in Fig. 34(*k*), which shows an all-wave balanced system having separate aerials for short and for medium and long waves. The disadvantage of such systems is that they are definitely directional when horizontal spacing is used, while if vertical spacing is used it is extremely difficult if not impossible to ensure that the system is balanced at all wavelengths.

In all systems it is necessary to provide a conductive path between the aerial and earth to prevent the accumulation of static charges on the aerial, and where the connection is not inherent in the circuit it can be provided by centre-tapping one or more of the transformer windings.

The primaries and secondaries of the transformers should be screened from one another, or other appropriate measures should be adopted in order that the balanced windings shall have equal capacitances to earth, and to avoid capacitance coupling between the primary and secondary windings. The transformer as a whole should preferably be screened, particularly in the case of the set transformer. Where interference is severe the receiver input lead should be screened also to within about an inch of the set, the screen being effectively earthed.

In the case of flats, where individual outdoor aerials are impossible, the best solution appears to lie in the provision of a single efficient aerial at the top of the building associated with a wide-band amplifier to the output of which a feeder is connected which is brought into every flat. The amplifier must be free from intermodulation effects, and provision must be made—by the insertion of a suitable impedance in the supply to each flat, or other means—to prevent as far as possible adjustments at the listener's receiver affecting other users of the service.

(ii) Type of receiver.

The increasing use of more sensitive receivers during the last few years has tended to accentuate the trouble due to interference, as the highly sensitive types now available render an efficient outdoor aerial unnecessary, with the result that there has been a tendency to use indoor aerials with their inherently higher coupling to the mains. At the same time the high sensitivity now available is a temptation to listeners to attempt reception of distant stations of low field-strength and subject to a higher degree of interference than the stronger local stations. The increase of interference perception due to the higher sensitivity has been offset to some extent by the improved selectivity of modern receivers, which, with their much narrower band of acceptance, reject much of the noise which would enter and affect the older and less selective type of receiver. In this connection, devices which allow the selectivity of the receiver to be increased when working on high gain are a valuable improvement. This improvement cannot be pressed beyond the limit where quality suffers.

The use of automatic gain control without a muting device for quenching the receiver in the absence of a carrier tends to give a false impression of interference, as the gain of the receiver rises to an excessive degree when it is being tuned between stations, with a corresponding increase in noise.

(iii) Mains filters.

The use of filters at the point of entry of supply mains to the listener's premises is a valuable aid towards the prevention of interference from plant outside the building connected to the mains, particularly in those cases where the level of interference or numbers of persons affected would not justify suppression at the source. Such filters normally consist of two condensers of capacitance 0·5–2·0 μF. One condenser is connected across the two supply mains, and the other condenser is connected from the neutral main to earth. Cases arise in which condensers alone suppress the interference to an inadequate degree; in such cases choke coils are used in addition. The condensers may be placed either on the street side of the chokes or vice versa, depending on the relative impedances of the street mains and the house mains. The best position can be found most easily by trial.

(c) Specifications and Procedure in Dealing with Interference

During the past few years a number of British Standard Specifications have dealt partially or wholly with radio interference. The first specification of the kind was B.S.S. No. 505—1933* [" Road Traffic Control (Electric) Light Signals "], which contained a clause requiring such equipment to be free from interference and an appendix dealing with electrical interference. B.S.S. No. 613—1935 (" Components for Radio-Interference Suppression Devices ") contains schedules specifying the recommended value of condensers, inductors, and resistors for different types of machines and appliances. B.S.S. No. 727—1937 (" The Characteristics and Performance of Apparatus for the Measurement of Radio Interference ") has already been referred to in Section (1). B.S.S. No. 800—1937 (" Standard Method for the Characteristics and Performance of Apparatus for the Measurement of Radio Interference ") defines the conditions for the issue of the Radio-Interference-Free mark.

Further specifications are in hand dealing with ignition systems, electromedical apparatus, trolley-buses and tramways, and radio receiving installations. B.S.S. No. 727 is in course of revision with a view to its extension to metre waves.

The procedure in Great Britain and Northern Ireland in dealing with interference is that all complaints, whether received by the B.B.C. or by the Post Office, are handed over to and investigated by the Post Office Engineering Department. Fig. 35 shows the number of complaints dealt with since 1929. The number of complaints dealt with in 1936 was 44 000, and the curve continues to rise. This increase in the number of complaints may be attributed to the increase in the number of listeners, the increase in the use of electrical appliances, and the growing knowledge of the public of the existence of the investigation service. In spite of the fact that the Post Office has given at exhibitions a certain amount of publicity to the work it is doing, it is believed that a very large number of listeners are still unaware of the service and are suffering in silence in conditions where it would be possible to provide relief. From information at the disposal of the Post Office it is believed that a considerable number of people, particularly those with houses in close proximity to electric railways, trolley-buses, and tramways, do not purchase licences or do not renew them because of their knowledge of the interference likely to be encountered; and that a large number of listeners tolerate more or less severe interference as they are aware that neighbours have complained or because they are too apathetic to register a complaint.

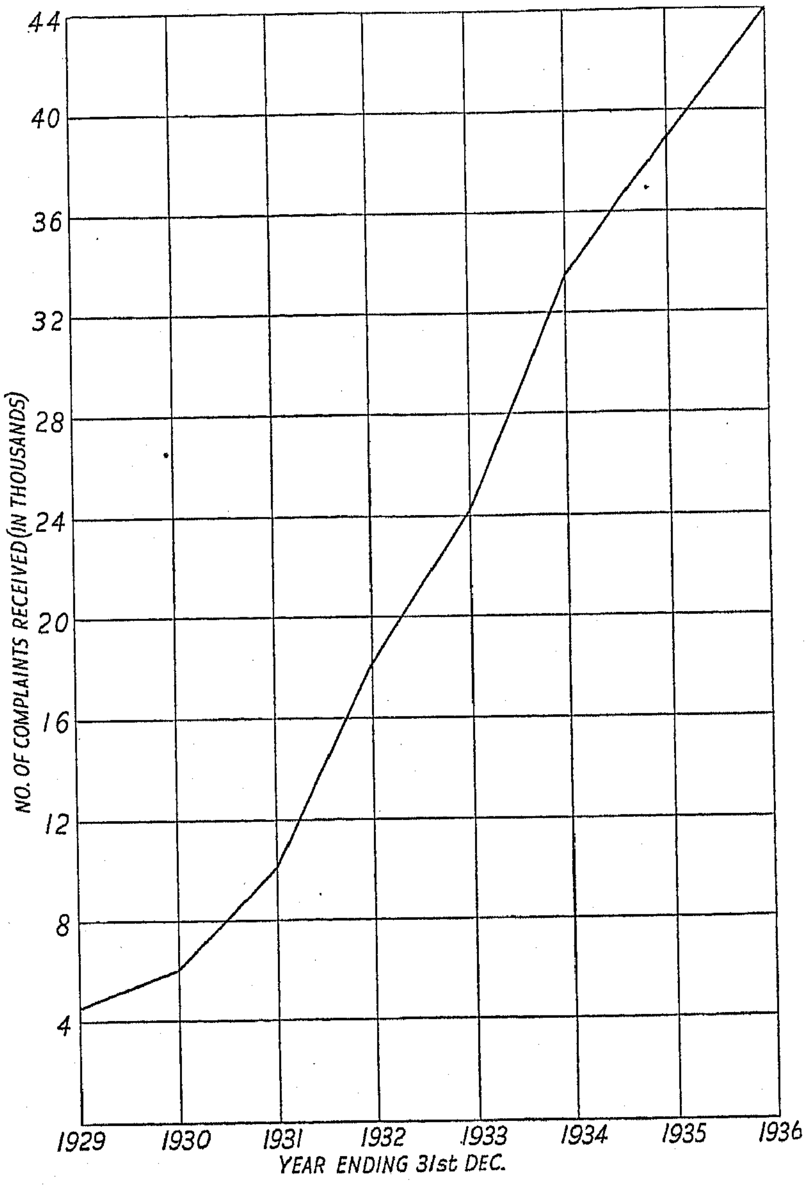

Fig. 35.—Interference complaints received by G.P.O.

In 1934 an analysis was made of 1 000 cases of interference, taken at random from all parts of the country, in order to ascertain the distribution of interfering sources, with the following result:—

Type of plant	Percentage of total
Motors	50
Trams, trolley-buses, and electric railways ..	10
Flashing signs	5
Neon signs	4
Medical apparatus	4
Rotary rectifiers	3
Static rectifiers	3
Overhead power lines	2
Overhead telegraph and telephone lines ..	1
Underground power line	1
Railway signals	0·5
Traffic signals	0·5
Miscellaneous plant and apparatus	9
Faults at listeners' premises	7

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* Current edition is B.S.S. No. 505—1937.

The number of equivalent full-time Post Office staff employed in Great Britain and Northern Ireland on broadcast interference work is now of the order of 250. In order to improve their efficiency and widen their experience, special courses of instruction are held, by means of lectures and special demonstrations. The equipment available to this staff comprises 110 small motor vans, each supplied with an interference locator, portable wireless receiving set, screened leads, and a miscellaneous assortment of condenser filter units and inductors for experimental demonstration purposes. At the present time the investigations are carried out on the medium- and long-wave broadcast bands, but additional equipment is being provided to enable such work to be extended to the television band in the London area. The portable wireless receiving sets are of normal broadcast type, with self-contained batteries. Both h.f. amplification and the superheterodyne type are used. The sets are provided with internal frame aerials, and in addition have terminals for external aerial and earth connections. Jacks are also provided for headphone reception as an alternative to loud-speaker. The use of these sets enables tests to be made to verify that interference is actually present, and that the trouble is not due to faults on the complainant's apparatus. More recently a portable interference-locator has been developed by the Department, and it is the intention eventually to replace the majority of commercial types of portable receivers in use for interference work with this apparatus.

It is gratifying to record the co-operation which has been afforded by the parties involved. All branches of the electrical industry, the motor industry, the electro-medical trades, and the medical profession, have assisted in the preparation of and the necessary studies for the British Standard Specification relating to their equipment. The B.B.C., G.P.O., and Radio Manufacturers' Association have made special subventions to the E.R.A. in support of its work in this direction. Finally, in general, members of the public have willingly granted facilities and have followed the recommendations of the G.P.O. in specific cases investigated.

(4) ACKNOWLEDGMENTS

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APPENDIX

Theory of Signal/Noise Ratio in Ideal Receivers

For the purpose of the present analysis it is assumed that the h.f. and l.f. amplifiers and associated equipment amplify uniformly all the components which lie within the frequency bands which they select. If this is the case, then the constant amplification factor may be omitted and it is only necessary to consider the effect of rectification and of the frequency-selection mentioned.

(1) Single interfering frequency.

The input voltage may be expressed as:—

$$V = P(1 + M \cos p_0 t) \cos (\omega + p)t + V_1 \cos \omega t \quad (1)$$

where

P = amplitude of the broadcast carrier,
M = modulation depth,
$p_0/(2\pi)$ = modulation frequency,
V_1 = interfering voltage,
and $p/(2\pi)$ = frequency-difference between interfering voltage and carrier.

In the simplest form of ideal linear rectification all the half-waves of voltage of a given sign are faithfully transmitted while the half-waves of opposite sign are suppressed. Mumford and Stanesby* have represented this action very conveniently by the unit operator function:—

$$f(t) = \tfrac{1}{2} + P_1 \cos \omega' t + P_3 \cos 3\,\omega' t + \ldots \quad (2)$$

where $$P_n = (-1)^{\frac{1}{2}(n-1)} \frac{2}{n\pi}$$

n being an odd integer.

This function is alternatively $+1$ and zero, with a mean period of $2\pi/\omega'$. Mumford and Stanesby regarded ω' as constant. This assumption is sufficiently accurate for the deduction of the main principle of interference measurement, but for greater certainty it is desirable at first to extend the analysis, since a combination of components of different frequencies does not necessarily give rise to a function of constant period and therefore ω' will usually vary in practice, giving a species of "frequency modulation." The two components of V can be regarded as vectors of angle ωt and $(\omega + p)t$, and the angle of the resultant is $\omega' t$, where

$$\omega' t = \omega t + \phi$$

and $$\phi = \arctan \frac{P(1 + M \cos p_0 t) \sin pt}{V_1 + P(1 + M \cos p_0 t) \cos pt} \quad . \quad . \quad (3)$$

Forming the product $Vf(t)$, we have

Rectified voltage

$$= P(1 + M \cos p_0 t) \cos (\omega + p)t \left[\tfrac{1}{2} + P_1 \cos (\omega t + \phi) + \ldots\right] + V_1 \cos \omega t \left[\tfrac{1}{2} + P_1 \cos (\omega t + \phi) \ldots\right] \quad . \quad . \quad . \quad (4)$$

This expression gives a series of products which can be split up into summation and difference tones. Only those of low frequency will be accepted by the l.f. amplifier, so that terms involving ω or its multiples may be neglected. Thus the output due to V is given by:—

$$2V = PP_1(1 + M \cos p_0 t) \cos (pt - \phi) + V_1 P_1 \cos \phi \,.\,. \quad (5)$$

If $M = 0$, then

$$\frac{2V}{P_1} = (P^2 + V_1^2 + 2V_1 P \cos pt)^{\frac{1}{2}} \quad . \quad . \quad . \quad (5.1)$$

If M is not zero,

$$\frac{2V}{P_1} = P \cos (pt - \phi) + V_1 \cos \phi + PM \cos p_0 t \cos (pt - \phi) \quad . \quad . \quad . \quad (5.2)$$

The presence of V_1 multiplies the modulation term $(M \cos p_0 t)$ by a factor $\cos (pt - \phi)$, where

$$\cos (pt - \phi) = \frac{V_1 \cos pt + P(1 + M \cos p_0 t)}{\left[V_1^2 + 2PV_1(1 + M \cos p_0 t) \cos pt + P^2(1 + M \cos p_0 t)^2\right]^{\frac{1}{2}}}$$

$$= \frac{V_1 \cos pt + P(1 + M \cos p_0 t)}{\left[V_1^2 + P^2(1 + M \cos p_0 t)^2\right]^{\frac{1}{2}}} \cdot \left[1 + \frac{2V_1 P(1 + M \cos p_0 t) \cos pt}{V_1^2 + P^2(1 + M \cos p_0 t)^2}\right]^{-\frac{1}{2}}$$

Neglecting M,

$$\cos (pt - \phi) = \frac{V_1 \cos pt + P}{(V_1^2 + P^2)^{\frac{1}{2}}} \left\{1 + \left[\frac{2V_1 P \cos pt}{V_1^2 + P^2}\right]^{-\frac{1}{2}}\right\}$$

$$= \frac{V_1 \cos pt + P}{(V_1^2 + P^2)^{\frac{1}{2}}} \left[1 - \frac{V_1 P \cos pt}{V_1^2 + P^2} + \ldots\right]$$

$$= \frac{P}{(V_1^2 + P^2)^{\frac{1}{2}}} + \frac{V_1 \cos pt}{(V_1^2 + P^2)^{\frac{1}{2}}} - \frac{V_1 P^2 \cos pt}{(V_1^2 + P^2)^{\frac{3}{2}}} - \frac{V_1^2 P \cos^2 pt}{(V_1^2 + P^2)^{\frac{3}{2}}}$$

$$= \frac{P}{(V_1^2 + P^2)^{\frac{1}{2}}} - \frac{V_1^2 P}{2(V_1^2 + P^2)^{\frac{3}{2}}} + \cos pt \left[\frac{V_1}{(V_1^2 + P^2)^{\frac{1}{2}}} - \frac{V_1 P^2}{(V^2 + P^2)^{\frac{3}{2}}}\right]$$

and terms in $2pt$, $3pt$, etc.

If P^2 is neglected as compared with V_1^2 in the first two terms,

$$\cos (pt - \phi) = \frac{P}{2V_1} + \cos pt \left\{\frac{1}{\left[1 + (P/V_1)^2\right]^{\frac{1}{2}}} - \frac{(P/V_1)^2}{\left[1 + (P/V_1)^2\right]^{\frac{3}{2}}}\right\} \quad (6)$$

Since ϕ depends on V_1, P, and M, the "normal" terms $P \cos (pt - \phi)$ and $V_1 \cos \phi$ give rise to modulation products. This form of distortion is a second-order effect in most instances, and is neglected in the classical analysis. It is also neglected here, where we have only followed the transformation of the normal modulation tone from the point of view of its amplitude, without considering its distortion in any detail.

The first term represents the apparent decrease of modulation caused by the presence of another carrier, the "demodulation effect."* The second term represents the intermodulation between V_1 and the sidebands of P, which is only affected to the second order by the carrier of P.

If P is large compared with V_1, ϕ tends to (pt), i.e. ω' corresponds to the carrier frequency, and

$$\frac{2V}{P} = PM \cos p_0 t + V_1 \cos pt - \frac{3}{4}\left(\frac{V_1}{P}\right) V_1 \cos 2pt + \ldots \quad (7)$$

$$= PM \cos p_0 t + V_1 \cos pt \text{ (approx.)}$$

Thus the interfering voltage V_1 appears undisturbed in the output but with its frequency changed to the difference between the original value and the carrier frequency. The signal/noise ratio is thus MP/V_1 if both are observed in a similar manner.

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* A. H. Mumford and H. Stanesby: Radio Report No. 300 (P.O. Engineering Department).

* For the classical analysis of this effect, see E. V. Appleton: *Wireless Engineer*, 1932, vol. 9, p. 136; and F. M. Colebrook: *Wireless World*, 1931, vol. 28, p. 560.

(2) Several interfering frequencies.

For convenience, let the broadcast signal be

$$P(1 + M\cos p_0 t)\cos\omega t$$

Then $\omega' t$ is, as before, $(\omega t + \phi)$, where

$$\phi = \arc\tan\frac{\Sigma a_k \sin(p_k t + \alpha_k)}{P(1 + M\cos pt) + \Sigma a_k\cos(p_k t + \alpha_k)} \quad . \quad (8)$$

provided the summation includes all the interfering voltages like

$$a_k\cos\left[(\omega + p_k)t + \alpha_k\right]$$

which lie in the band accepted by the receiver.

Equation (5) now becomes:—

$$\frac{2V}{P_1} = P(1+M\cos p_0 t)\cos\phi + \Sigma a_k\cos(p_k t + \alpha_k - \phi) \quad (9.1)$$

If P is large compared with the interfering voltages, e.g. if the ratio is greater than 20 or 30 db.,

$$\sin\phi = \frac{\Sigma a_k\cos(p_k t + \alpha_k)}{P(1 + M\cos p_0 t)},\ \text{and}\ \cos\phi = 1 \quad . \quad . \quad . \quad (9.2)^*$$

Thus

$$\frac{2V}{P_1} = PM\cos p_0 t + \Sigma a_k\cos(p_k t + \alpha_k) + a_k\sin(p_k t + \alpha_k)\frac{\Sigma a_k\sin(p_k t + \alpha_k)}{P(1 + M\cos p_0 t)}$$

Considering the summation term, it is clear that the amplitude of each component is only affected to the second order, since $\Sigma a_k/P$ is small. If the first order of small quantities is neglected, then

$$\frac{2V}{P_1} = MP\cos p_0 t + \Sigma a_k\cos(p_k t + \alpha_k) \quad . \quad (9.3)$$

The signal/noise ratio is the ratio of the first to the second of these terms. If the noise is removed, then the signal is correctly measured; and if the modulation is removed, there remains

$$\Sigma a_k\cos(p_k t + \alpha)$$

which gives the noise correctly. Again, in the absence of a carrier, the h.f. interference is:—

$$\Sigma a_k\cos\left[(\omega + p_k)t + \alpha_k\right]$$

and it is therefore immaterial whether the h.f. interference in absence of a carrier is measured or the l.f. noise with a carrier is measured, provided a similar voltmeter is employed and provided that $1/\omega$ is much less than $1/p_k$ and much less than the charging time-constant of the voltmeter, and also that

$$\Sigma a_k\cos(p_k t + \alpha_k)$$

is continuous near its maximum value. If an acoustic filter is incorporated in the l.f. voltmeter, then the selectivity curve of the h.f. voltmeter must be correspondingly modified, but the time-constants should be the same. It is essential that the h.f. band should be correctly selected in the h.f. method, but in the l.f. method some of the selection may be made after rectification without important error, although it is desirable that the selectivity should be attained on the h.f. side. If the spectrum were continuous and uniform, all the a's would be equal and the p's would be in arithmetic progression. If, further, the phases were equal or completely incoherent in time, the magnitude of the interference would be proportional to the band width or integral of the selectivity curve. Since these conditions usually hold, it is possible to correct for a non-standard band-width by a simple multiplying factor.

If the interfering voltage is not small compared with the carrier, the output can be obtained by expanding (9.1), and a complicated form of distortion results. It has already been noted that, to a first approximation, the amplitude of the noise terms is unaffected, a phase variation being superposed, so that the error at first increases only slowly with the relative magnitude of the interference.

If there is no carrier, P is zero and only the second term of (9.1) remains. From (8) it is seen that ϕ will pass through the various values $(p_k t + \alpha_k)$ in turn, so that the resulting noise will contain all the intermodulation tones corresponding to the value of $(p_n - p_k)$. Thus the character of the noise is changed from that when the carrier is present. However, if the spectrum is continuous, the same frequency band will be covered, but the distribution will be different, since, other things being equal, the number of intermodulation products of a given frequency increases proportionately to the difference between the upper frequency-limit of the l.f. band and the frequency under consideration. Thus a filter giving such a weighting might be an advantage.

Since the standard voltmeter is largely a peak voltmeter, it is of interest to consider the peak value of the interference in absence of a carrier. From (9.1),

$$\frac{2V}{P_1} = \Sigma a_k\cos(p_k t + \alpha_k - \phi) \quad . \quad . \quad (10)$$

$$\phi = \arc\tan\frac{\Sigma a_k\sin(p_k t + \alpha_k)}{\Sigma a_k\cos(p_k t + \alpha_k)}$$

Choose the origin for time such that

$$\sum_0^{n_0} a_k\sin\alpha_k = 0 \quad . \quad . \quad . \quad . \quad (11)$$

That is,

$$\sum_0^{n_0} x_k = 0$$

over the range selected.

Let k be proportional to the frequency of the component, i.e. p_k is proportional to k, a hypothesis which is usually legitimate. Then, by analogy with the integral of a product,

$$\sum_0^{n_0} kx_k = n_0\sum_0^{n_0} x_k - \sum_{k=0}^{k=n_0}\sum_0^{k} x_k = -\sum\sum x_k$$

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* Neglecting the constant term.

from (11), provided n_0 is large. Now if x_k is distributed at random, then $\Sigma\Sigma x_k$ has a most probable value of zero. Thus

$$\Sigma k x_k \simeq 0 \simeq \Sigma p_k a_k \sin x_k \quad . \quad . \quad . \quad (12)$$

Now as $t \rightarrow 0$, $\phi \rightarrow 0$, from (11). Thus

$$\frac{-2V}{P_1} \rightarrow \Sigma a_k \cos \alpha_k$$

But, since (12) is a most probable condition,

$$\left[\frac{\partial}{\partial t}\Sigma a_k \cos(p_k t + \alpha_k)\right]_{\rightarrow 0} \simeq 0 \quad . \quad (12.1)$$

That is,

$$\Sigma a_k \cos(p_k t + \alpha_k)$$

has probably its maximum value, while it can similarly be deduced that in the same conditions (10) has an equal probability of a maximum value. It may therefore be said that, provided the number of lines in the noise spectrum is large and the distribution of phases and amplitudes is random, the most probable condition is that the noise either in presence or in absence of a carrier will have the same peak value and is therefore likely to give the same reading on the voltmeter employed. It is not easy to place an absolute value on this probability, but the deduction as regards this general case is reinforced by the experimental fact that the carrier makes little difference in a large number of instances where the interference is of a more or less continuous nature. If the phases and amplitudes are equal a similar conclusion can be reached in a simpler manner.

Consider the simple but frequent case of a square pulse where all the components near a given frequency ω have the same amplitude and phase, while $p_k = k_p$, since the interval between successive components is constant. Thus

$$\phi = \text{arc tan} \frac{\Sigma \sin kp_0 t}{\Sigma \cos kp_0 t} \quad . \quad . \quad . \quad (13.1)$$

whence

$$\begin{aligned} \frac{2V}{P_1} &= a\left[(\Sigma \cos kp_0 t)^2 + (\Sigma \sin kp_0 t)^2\right]^{\frac{1}{2}} \\ &= a \text{ mod.} \Sigma e^{jkp_0 t} \\ &= a \text{ mod.} \left(e^{jp_0 t} \frac{1 - e^{jnp_0 t}}{1 - e^{jp_0 t}}\right) \\ &= a\left(\frac{1 - \cos n^{p_0 t}}{1 - \cos p_0 t}\right)^{\frac{1}{2}} \quad . \quad . \quad . \quad (13.2) \end{aligned}$$

where n is the number of components selected in the band width.

The maximum value of $2V/P_1$ is given by $t = 0$ and is found by determining the limit of (13.2), which is na. Similarly, the maximum value of

$$\Sigma a \cos(pk + \omega)t$$

is na, so that the l.f. voltmeter without a carrier will give an approximately correct measure of the interference.

It is clear that, in general, errors will arise from the omission of the carrier. These errors may be of either sign, but in practice the curvature of the actual detector characteristic exercises a general determining influence, such that with the carrier present the measured noise is greater, since the rectifier is operating over a straighter and more sensitive portion of its characteristic.

DISCUSSION BEFORE THE WIRELESS SECTION, 6TH APRIL, 1938

Mr. A. H. Bennett: Some impatience has been expressed in the radio, the technical, and the semi-technical Press on the subject of radio interference. Some people have thought that the electrical industry and the radio industry should have cured interference long ago; but it is a very complex problem, and many interests have to be brought into line before a solution can be reached. Great Britain is as near the desired result at the present time as probably any country in Europe.

I am glad to be able to report from several sources that since the trolley-buses were put into service in North London, approximately 3 weeks ago, the amount of radio interference has been greatly reduced. There has been a certain amount of interference caused by the tramways, but the trolley-buses are relatively free.

Mr. F. M. Colebrook: I notice that the authors use the term "mains attenuation" as an alternative to "coupling factor."* "Mains attenuation" has a fairly definite meaning, quite different from that implied in this paper, and I therefore prefer the use of the term "coupling factor." I hope that this term will be employed in future if either of the terms comes into common use.

Mr. R. T. B. Wynn: I cannot believe that there can be real contention on the question of who is responsible for silencing interference. The broadcasting service must do its part in trying to make things a little easier for the other sections of the electrical industry, and vice versa. It is frequently stated that a great many listeners have decided to come on to the public supply mains because they want to use all-mains receivers. If they do, they become a market for electrical appliances, which they will expect to be interference-free. I should like to quote two figures which may be of interest. Sir Felix Pole said recently that approximately 35 % of the householders in this country are not yet connected to public supply mains. The number of homes without wireless licences is about 33 %. There is a similarity between these two figures which points to a rather obvious moral.

It may cause great confusion if the B.S.I. "interference-free" mark becomes optional. If Mrs. A is very keen on listening and wants to buy a sewing machine, she may think that she could not use it when listening, and therefore need not pay the extra half-guinea for the two condensers. But she would be very angry if Mrs. B were to think the same about her violet-ray machine.

On page 210 of the paper it is stated that "the signal

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* Corrected for the *Journal*.

(programme)-to-noise ratio so determined as corresponding to the limit of toleration was then found to be substantially independent of the observer, type of programme, and type of noise. . . ." It is agreed that there is only one way to measure the strength of the wanted signal and that is to measure the peaks, because mean modulation in a broadcasting programme is so indefinite. Attempts to measure mean modulation have led into all sorts of difficulties, because, for example, we do not know whether to include the time when the orchestra is having a rest between two movements. There are other and less obvious factors to be considered, including the reverberation of the studio. Certain programme producers actually use reverberation to get more of a sense of climax without exceeding the peak modulation levels.

I believe that the degree of tolerable interference is dependent on the type of programme, and must be set at a ratio which will satisfy the most difficult items. There are certain programmes, such as a full symphony orchestra, which demand all the dynamic range that can be given to them. This must mean that the peaks occur less frequently, and therefore a given ratio of interference based on peak modulation must sound worse. The problem is made particularly serious by the type of audience which is attracted by such programmes. Symphony-concert audiences go through long periods of very low modulation, during which they are listening very intently indeed, even when the orchestra is not playing, in a pause between two movements. I should like the authors to give a little further explanation for their statement that an acceptable interference ratio is independent of the type of programme. I hope that they mean the ratio must be set at a level which will satisfy all types of programme.

Mr. T. H. Kinman: I find that it takes 6–8 weeks to train an operator in the technique of using the interference measuring set shown in Fig. 2. It is not simply a matter of reading a meter or turning a knob, but of interpreting the measurement, and the ability to do this can come only with experience. The point that I want to make is that if it becomes necessary to equip every factory with such precision apparatus, there may be difficulties in finding suitable personnel to operate the testing equipment.

I am surprised that the authors do not mention in Section (2) such potential sources of interference as the radio installation itself, domestic lighting and power circuits, earthing conductors (when used), terminal and fuse boards, lamps and lamp fittings, and plugs and sockets. All these items are frequent sources of interference of a most elusive and annoying type, and I think that much could be done to remove such causes of interference by the more strict observance of the regulations, including the I.E.E. Wiring Regulations. Some radio-interference statistics recently published in America showed that in a certain State no fewer than two-thirds of the complaints arose from items on the listeners' own premises, and many of the items were of the sort to which I have referred.

The details given on page 244 regarding the distribution of interfering sources might have been made somewhat more informative, and certainly more up-to-date; for instance, the term "motors" covers a very wide range, from about 50 watts upwards.

The authors treat very lightly the question of what can be done to correct the appliance itself before additional suppression is added. Practically every type of apparatus could be improved by some change in design or assembly, whereby any additional suppression required would be rendered easier and less costly. Even more important is the need for investigation into the many causes of the wide variations observed now in terminal interfering voltages on different samples of apparatus of a particular type, which makes it difficult to suppress to a common denominator. I say this because the authors imply that certain rules can now be used for dealing with the smaller items.

Mr. E. M. Lee: I want to refer to the figure of 500 μV that is laid down in Clause 2(*a*) of B.S.S. No. 800—1937. I have heard it suggested, particularly abroad, that we are being rather too ambitious; for the long wave band where it is most difficult and most expensive to apply suppression, there is a suggestion on the Continent that the figure should be 1 500 μV up to a maximum wavelength of 2 000 metres. There is not so much difference between our figure and the Continental suggestion as may at first appear, because our 500 μV applies up to a maximum wavelength of 1 500 metres, and is therefore equivalent to about 1 000 μV at 2 000 metres.

We should give the present standard a good trial before there is any question of reconsidering it or allowing any tolerance. One of the authors' slides showed that 64 % of earthed machines—i.e. machines in the most difficult condition—can be suppressed with condensers alone to 200 μV asymmetric voltage. With such a standard of suppression the worst of the machines with their worst adjustment after some years of wear would still give a figure of about 500 μV. I think that those prescribing suppression should purposely put machines out of adjustment, getting the brushes in the worst possible condition so as to imitate the state of affairs when the machine is rather old, and then try to keep below 500 μV.

We have the difficulty that about one-third of the machines when earthed will not be suppressible with condensers alone. Some electrical designers, when confronted with B.S.S. No. 800—1937 as recently as November last, tried the experiment of fitting condensers and did not immediately get the required suppression. I would point out that suppression engineers who have been doing nothing else for 4–5 years sometimes have to work for days to find a way of getting down to the required level, but once the trick has been discovered for a particular machine it holds good for that machine. A skilled suppression engineer would not feel that his task was complete unless he turned out his sample machine with a level of 50 or 100 μV over a wave band far beyond the coverage of the present Specification; for, it is usually very little more difficult to extend the range down to the short and ultra-short wave bands.

To make suppression most successful one must reduce the length of the condenser leads to $\frac{1}{4}$ in. if possible; often a reduction from $\frac{1}{2}$ in. to $\frac{1}{4}$ in. avoids the use of chokes. The use of a suppressor in a flexible lead is only a temporary expedient to be employed until the manufacturer finds it convenient to re-design and put his condensers where they ought to be, namely inside the casing.

Mr. G. A. Struthers: One curious feature of the authors' results is the great amount of difference in the levels of interference produced by different machines. For example, with vacuum cleaners of more or less the same horse-power and design we get interference varying from 2 mV to 100 mV. Some of the more expensive machines give far more interference than the cheaper ones; actually, the most expensive vacuum cleaner which the Post Office ever tested had the highest level of interference.

It would be worth while to investigate the reason for the disparity which exists between machines of different makes, and also between individual machines of the same design and made in the same way. If we confine our attention to a single type of motor, there seem four possible reasons for this disparity; namely the number of commutator segments, the electrical balance of the windings, the mechanical balance of the machine, and, in the larger machines at least, possibly the question of lubrication. Many portable vacuum cleaners are already too heavy, and if anything could be done—e.g. by changing the type of motor or improving the design—to avoid the need for suppressors, it would be of tremendous help to the people who have to use the machines.

Mr. E. A. Watson: I should like to deal with the question of ignition interference from motor-cars.

Such ignition interference is primarily a question of interference with ultra-short-wave systems. The motor-car does not interfere with ordinary broadcast reception, and I do not think the paper is intended to deal with the question of the suppression of the ignition equipment of the motor-car for the sake of any radio receiver which is carried on the car itself. The suppression of the interference which the motor-car may cause to short-wave systems in the neighbourhood will therefore involve fitting suppression devices to every motor-car in the country, a very difficult matter involving, I should estimate, a capital outlay of about £2 000 000.

The authors mention various methods of suppression. The ordinary method of braiding and bonding which is used on aeroplanes is particularly difficult to apply to a motor-car. On an aeroplane, which normally runs on full throttle, one operates with relatively short plug-gaps and as the plug temperatures are invariably high in addition the plug voltages are relatively low, of the order of 5 kV or less; but on a motor-car, which has to work over wider ranges of throttle opening with larger plug-gaps and in general lower plug temperatures, considerably higher voltages (some twice as great or even more) are necessary. These higher voltages may lead to a good deal of insulation leakage and, still worse, to insulation deterioration and breakdown, while the increased energy required by the capacitance of the braiding may lead to troubles of various sorts in both primary and secondary circuits. Fortunately, as the authors point out, in most cases complete suppression will not be necessary, and in many cases resistances in the leads will suffice, but it is not yet certain whether resistances can be used on all motor vehicles without their having some effect on the performance of the engine; and the public, while being quite willing to go to some trouble and expense to avoid interference on short-wave reception, may not be inclined to install equipment which adversely affects the running of their motor-cars.

A great measure of improvement can be obtained if the coil is mounted on the engine, but unfortunately the present-day motor-car manufacturer mounts his engine on rubber to make it sweet-running and then takes advantage of the rubber mounting by making the engine "rough," so that it vibrates to a serious extent. It is rather unfair to expect a delicate apparatus, which has to deliver a voltage up to 10 000 volts, to be mounted on a crankcase at temperatures up to 100° C. and to be shaken about violently and yet to have an indefinite life without being given any attention. While, therefore, the automobile engineer is cognizant of the requirements and of the need for avoiding unnecessary interference with short-wave reception, I would wish to make it clear that the solution of the problem is not in all cases quite so simple as would appear at first from the paper.

Mr. E. C. S. Megaw: I should first like to make some remarks on the subject of motor-car ignition interference. This interference reaches a maximum at frequencies of the order of 25–50 Mc./sec., i.e. in the neighbourhood of the present television bands. We have found that the general level of interfering field-strength from motor-car ignition systems at frequencies of 300–600 Mc./sec. is much lower, probably 20–30 db. below the level at frequencies of about 50 Mc./sec. The dependence of ignition interference level on frequency is thus a factor to be considered in planning future short-range radio services.

The rest of my contribution is concerned with the interference from electrical therapy machines in the ultra-short wave band. The use of ultra-short waves for such purposes is increasing fairly rapidly, and for reasons which are partly medical and partly economic the wavelength of 6 m. has become the prevailing standard for much of this work. So far as spark apparatus is concerned there is very little to be done except to adopt the most complete screening possible. In the case of valve apparatus, which in this field, as in the field of communication, is gradually displacing spark apparatus, the problem is essentially simpler because the radiation is confined to relatively narrow frequency bands. Table A gives some data on the effects of different kinds of screening on the radiation from a 6-m. valve therapy machine. In one test a complete copper-sheet-screened room was tried in which the electrical continuity was as perfect as possible; special methods were used to seal the door, and there was complete filtering of the mains supply leads. Two methods were used—a completely screened isolation transformer and a special condenser filter. The latter, employing a pair of cylindrical condensers of very low residual inductance, was quite as effective as the isolation transformer, and much cheaper. Another device which was tried was to place a strip of copper foil round the load circuit of the machine, with the idea of limiting and partially short-circuiting the radiation field. Although we tried several variants of that scheme, it was not very successful; but it might be useful for reinforcing the other methods of screening. Some rough measurements which we made of the field strength from a 6-m. electrotherapy machine at a distance of the order of 10 yd. come within 10 db. of the values in Curve A of the authors' Fig. 13, which is a very plausible agreement. Even with the best possible kind of screening, and the best possible mains filtering, if the radiation from a valve machine of about 0·5 kW output lies within the television frequency band, and there is a television receiver in the same building, there will in almost all cases be appreciable interference. On the other hand, if the frequency of any valve machine is outside the television band, the problem of eliminating interference is enormously simplified and the solution is quite cheap; in some cases no screening at all is necessary if mains filtering is provided. We found that mains filtering is always essential. In this connection it would be interesting to know the frequency of the valve set used in the authors' demonstration.

In connection with the question of frequency spacing the properties of the radiated signal are of importance. Our measurements showed that very roughly the frequency band occupied by a 6-m. valve set working on unrectified a.c. anode supply is about ± 1 % from the carrier frequency. The attenuation to which that corresponds is probably between 40 and 60 db. The frequency-change of such machines on heating-up was in our sets less than 1 %. Changes due to alteration of the load circuit depend very greatly on the design of the machine, but need not exceed 1 % or 2 %, so that in order to avoid interference completely a frequency-spread of the order of a few per cent should be allowed.

We have now to face the situation in which a large number of diathermy machines are operated on a nominal frequency of 50 Mc./sec. (How nominal that figure sometimes is can be realized from a case which the authors quote where a so-called 6-m. set gave a fundamental frequency coinciding with the vision transmission from Alexandra Palace.) If we can agree that the manufacturers of such machines will confine their frequency to a specified band—a tentative figure of 48 to 52 Mc./sec. occurs to me as reasonable, but closer limits have already been adopted by the manufacturers with whom I am associated—and, on the other hand, if the designers of television services will regard such a band as occupied in making plans for future transmitters, interference between apparatus of this kind and television can be completely avoided. On the other hand, if no such arrangement is made, limiting by agreement the present legal rights of both parties, this kind of interference is likely to be a serious and increasing embarrassment to the development of television.

Table A

ATTENUATION OF RADIATED FIELD FROM A 6-M. 0·3-KW VALVE THERAPY MACHINE BY SEVERAL TYPES OF SCREENING ENCLOSURE: DISTANCE APPROXIMATELY 10 YD.

Type of screening	Attenuation (db.)	Mains filter
(1) Copper sheet, electrically continuous; door clamped to give contact all round	60	Completely screened isolation transformer, *or* special condenser filter (2 × 0·006 μF)
(2) As (1)	18	None
(3) " Fine " copper mesh (about 150 to the square inch), joints overlapped 1 in. and soldered at about 3-ft. intervals. Screening on door makes contact with main screen, though less perfectly than in Case (1)	30	Partially screened isolation transformer
(4) " Medium " copper mesh (about 12 to the square inch), joints not soldered, window not screened	14	None (receiver and oscillator supplied from same mains)
(5) As (4), but with window screened	25	None (receiver out of doors and supplied from separate source)
(6) Copper-foil strip 5 in. wide surrounding patient circuit and load, spaced about 6 in. from patient-circuit leads and earthed to metal frame of oscillator	7	Partially screened isolation transformer

Mr. W. Hill: In common with Mr. Kinman, I am very surprised at the figures for the distribution of interfering sources given on page 244, and particularly at the last two items. In my experience, house wiring is a much more common cause of interference than is there indicated. Do the authors include refrigerator motors under the general heading " motors "? I have found these motors among the most troublesome items to deal with.

With regard to what the authors say on page 217, have they had experience of the use of insulators on power lines with a metallized coating? These have been very successful in America. A comparison of insulators on a surface-resistance basis gives an indication as to which one is going to effect a cure of radio interference, and I recommend also the use of the high-frequency test. Many insulators have proved faulty on that test which can stand up to the ordinary dry and wet flashover test and yet be the cause of radio interference. For locating any faulty insulator we have found the Tebo stick most reliable and easy to operate. In one case which came to my notice recently, it was suggested that there was a faulty insulator on a 33-kV overhead line some 6 miles away from the wireless set which was giving trouble. The interfering voltage was supposed to be transmitted along the 33-kV line and through the transformer; but I did not believe this possible. I would ask the authors whether interference from a faulty insulator on a 33-kV line can be transmitted through the transformer. I should have thought that the impedance of the transformer would have been too high and would have choked any high-frequency operation. The fault was finally located on the wiring next door!

I should like to ask whether an increase in the current being transmitted on an e.h.t. power line does not cause an increase in the level of radio interference when this is present.

Mr. E. L. E. Pawley: It is astonishing how little is known about the efficiency of receiving aerials. One would have supposed that every elementary textbook would give a chart showing the effective height and the impedance to be expected from every shape and size of receiving aerial, whether indoor or outdoor, with or without screened down-leads; but that is not the case. We know just what it would cost us to give 1 db. more output from Droitwich, and the manufacturers know more or less what it would cost them to suppress the interference from a machine by 1 db.; but the listener has not the remotest idea what he must do to improve the signal/noise ratio by 1 db.

Mr. Watson spoke about motor-car ignition systems not affecting ordinary broadcast reception; he must have been referring to broadcast reception on medium and long waves, because interference of this type can be very severe on short-wave broadcast reception. As regards interference with television, the authors give some figures for the signal/noise ratio which has been found tolerable; but I would point out that the figure which can be tolerated depends very much on the nature of the picture, the type of interference, and the part of the picture which it affects.

There is still a great deal of work to be done in making more complete our information on interference at the higher frequencies.

The education of the listener and of the general public in the matter of radio interference is of vital importance. In particular, listeners must be encouraged to use efficient aerials, and landlords must be asked not to put unnecessary restrictions on the aerials which their tenants are allowed to erect.

With regard to Fig. 7, is it an accident that the maximum interference on the long wave band coincides precisely with the wavelength of Droitwich? The scale of ordinates gives levels in decibels above the level at 200 kc./sec., and, this being so, I should like to know why the lower curve does not intersect the zero line on the decibel scale.*

Fig. 8 gives some useful figures of interference at certain frequencies caused by trolley-buses, and I hope it will be possible one day to extend that information to other frequency bands.

On page 231 appears the statement: " For example, if in a given case the effective height of aerial was 1 m. and the mains attenuation was 37 db., a field of 1 mV per metre would be adequately protected from mains-borne radiation." Now if one refers to Fig. 25 to find out just what that " if " means, one finds that in less than 20 % of cases is the mains attenuation actually 37 db. or more; in all other cases it is less. Fig. 23 shows that the effective height is 1 m. or more in only 70 % of the cases, leaving 30 % which will get a weaker signal than the authors bargain for. Multiplying those two probabilities together, one finds that an alarming percentage of our 8 million listeners may still be in a position to complain about interference, even when the limit laid down in B.S.S. No. 800—1937 is everywhere complied with.

Mr. J. S. Forrest: I shall confine my remarks mainly to interference due to high-voltage overhead power lines. It is convenient to divide this type of interference into two main classes: (1) interference due to abnormal conditions in the power network; and (2) interference under normal operating conditions when the overhead lines and associated equipment are in perfect working order.

Abnormal interference is due to a fault on the power system causing a discharge which generates radio-frequency energy. The disturbance may be propagated throughout the power system, thus giving rise to widespread interference. In such cases, however, it is usually a simple matter to locate the fault and cure the interference.

The type of interference which occurs under normal operating conditions represents a much more serious problem. The cause is discharges on the line insulators, either on the surface of the porcelain or between the porcelain and the conductor or insulator hardware. There is at present no complete cure for this form of interference, and it must be recognized that there is a certain minimum level of interference which is inseparable from the operation of a high-voltage overhead-line system. In the case of lower-voltage (e.g. 33 kV) lines, which are usually insulated with pin-type insulators, it is possible by using special types of insulator to reduce the interference. The most successful type of insulator has a very low capacitance so that the energy in the spark discharges is reduced, with a consequent reduction in the interference. It is not possible at present to effect any further improvement in the performance of the chains of cap-and-pin insulators employed on the highest-voltage lines.

The characteristics of this type of interference are well illustrated by the data given in the paper. For example, from Fig. 9 it will be seen that the interference decreases as the wavelength decreases, and it may be mentioned that it is quite possible to listen to 20-m. American broadcasting within a very short distance of 132-kV transforming stations. Similarly, the results given in Fig. 10 are borne out by practical experience; it is usually found that, at distances of more than 50 yd. from the transmission line, interference with broadcasting is quite negligible. The results recorded in Table 4 are also borne out in practice, as complaints of interference are received only under humid weather conditions. Incidentally, the most prolific sources of complaints are the newly developed housing estates in which the houses are built under the power lines. It is considered that in such cases the builder should provide a screened aerial system for the benefit of the tenants.

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* Since corrected for the *Journal*.

In Fig. 11 it is indicated that the attenuation along the line is very small, and this suggests that the interference at mid-span would not differ appreciably from that at the tower. I should like to know whether the authors have made any measurements which confirm this.

With regard to mercury-arc rectifiers, large installations of rectifying equipment may give rise to interference with receivers connected to the a.c. mains associated with this equipment. The rectifiers distort the a.c. voltage wave, the distortion usually taking the form of the introduction of 5th and 7th harmonics, and these are propagated throughout the a.c. system associated with the rectifiers and affect radio sets which are not provided with particularly efficient smoothing equipment.

It is satisfactory to note that overhead power lines are responsible for only 2 % of the cases of interference referred to on page 244, particularly as there is no practicable means of suppression of interference of this type. Regarding the investigation of power-line interference, it has been found that the most convenient equipment consists of a standard car radio with an output meter and a switch to cut out the automatic volume control.

Captain J. McVicar: What is the relative effectiveness of galvanized-wire netting and expanded metal for screening purposes, and what size of mesh of expanded metal corresponds to the $\frac{1}{4}$-in. galvanized netting mentioned in the paper? I should also like to know how far double screening is effective, and whether there is an optimum distance between the two screens. From the point of view of construction it is easier to have the screens close together, say within $\frac{1}{2}$ in. of each other, and this practice should be adopted if it gives nearly as effective suppression as is obtained with the screens, say, 6 in. apart.

Finally, there are two practical points which I should like to put. In the case of the window with mesh in the glass, how are the ends of the wire which come out of the glass bonded? Does the means employed ensure that water cannot get to the mesh as it emerges from the glass? I have come across an interesting case of a piece of wired glass which was cut with a little mesh coming out from the ends; it was left out of doors for 3 weeks, and by the end of that time, at the point of emergence from the glass, the wire was corroded completely through. Have the authors any experience of the life of wire netting embedded in plaster? I imagine that with certain types of plaster there will be no wire netting left after a few years, unless precautions are taken.

Mr. F. R. W. Strafford: On page 228 the authors erroneously describe the mechanism of the effect whereby radio interference is produced by neon signs. This type of interference is nothing more than a relaxation oscillation of a type similar to that produced by ordinary gas-discharge devices associated with television time-bases. In the simplest time-base circuit a gas-discharge tube is connected across a capacitance, and this parallel arrangement is charged from a d.c. source of supply through a suitable resistance. Owing to the difference between the striking and extinction voltages of the gas, this resistance-capacitance combination will give rise to an oscillation whose wave-form is substantially sawtoothed in shape. Now the equivalent circuit of a neon sign is very similar to this arrangement excepting that the d.c. supply isreplaced by an a.c. source, the resistance is the equivalent series resistance of the high-voltage transformer, and the capacitance is the capacitance of the screened cable used for connecting the high-voltage transformer to the sign electrode.

In earlier laboratory investigations I failed to reproduce the interference solely because I had omitted to include the high-capacitance cable in the circuit. Unbonding of the shielding of this cable will in all cases reduce the interference to zero, unless, of course, the unbonding is accompanied by corona and other leakage effects.

Referring to the question of radiated interference, I feel a little worried about the constant use of the term "radiation" in connection with fields of interference which exist close to an interfering item. At a distance of less than 10 yd. from an interfering item the fields are obviously electric (scalar) or magnetic (vector) induction fields, and as such cannot be referred to in terms of volts per metre.

I am pleased to note the remark on page 243 that certain all-wave aerials with an unscreened twisted down-lead have no anti-interference properties on medium and long waves since their equivalent circuit is that of an ordinary T aerial. There have been so many instances of deceptive advertising claims in this respect that an authoritative announcement is welcome.

Mr. O. E. Keall (*communicated*): Referring to the theory of the signal/noise ratio in ideal receivers, given in the Appendix, it would appear that the application of the unit operator function $f(t)$ of equation (2) to a signal or signals determines the average value of the voltage of the unidirectional pulses of high-frequency energy obtained as a result of rectification. This may differ appreciably from the audio-frequency output voltage obtained in practice since the analysis takes no account of the detector load (or, alternatively, assumes a load without time-constant).

This omission may result in an estimate of signal and noise outputs differing appreciably from the actual output at the detector load terminals. For instance, the behaviour of the detector and its load at the beat frequency $p/(2\pi)$ is of some importance, for if the time-constant of the load (or the beat frequency, whether outside or inside the audio spectrum) is such that the detector cannot follow the beat-frequency variations, the noise will be much reduced and demodulation will not occur, with the result that actually a better signal/noise ratio is obtained than is indicated by the theory.

I would draw attention to an error in the last term of equation (4), which should read $P_1 \cos(\omega t + \phi)$.*

Mr. H. K. Robin (*communicated*): The paper contains no reference to the suppression of interference at the receiver, other than a brief description of all-wave antenna systems. The type of interference to which I refer is that which persists when the aerial has been removed and the receiver left at full sensitivity. I have been associated with some measurements on this type of interference, and would put forward the following description and results for the comments of the authors.

The mains interference voltage is considered to exist between both conductors and earth, i.e. the asymmetrical condition as described by the authors is assumed to exist. The method of entry into the receiver of this voltage is interesting and is not usually of the electromagnetic induction type but of a capacitive nature. Referring to Fig. A, the noise voltage V_1 is communicated to the chassis of the receiver by the various stray capacitances, together with a capacitance usually added with the intention of reducing mains noise; thus there now exists a second and somewhat smaller voltage V_2 between chassis and earth. If now the high-voltage end of the first tuned circuit has any small stray capacitance to earth, then a third voltage V_3 will appear by simple potentiometer effect across this tuned circuit, so producing a noise voltage which is amplified through the receiver and subsequently heard in the loud-speaker. If the stray capacitance from the high-voltage end of the first tuned circuit is made to chassis entirely, as opposed to earth, then no noise voltage will appear across the tuned circuit.

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* Since corrected for the *Journal*.

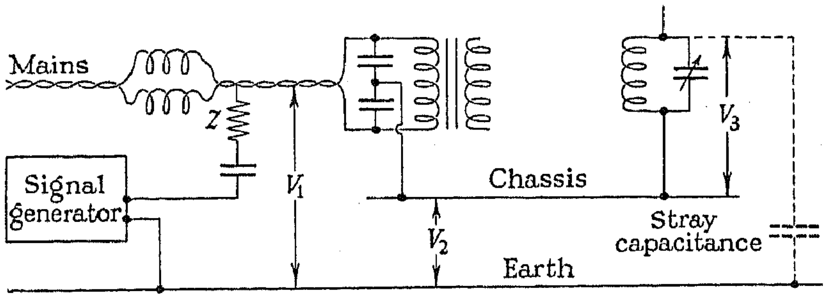

Fig. A

Various receivers were modified to fulfil this condition, and the following results were obtained: Receiver A, 30 db. improvement; Receiver B, 24 db. improvement; Receiver C, 40 db. improvement. In view of these results, we have tentatively instituted a new receiver measurement which we call the " sensitivity via the mains." It is rated as usual in microvolts for 50 mW audio-frequency output.

The method of measurement is, briefly, as follows: The signal generator is connected to the mains cord through an impedance (usually 100 ohms) simulating the source impedance of asymmetrical mains, and the input from the generator adjusted to give 50 mW audio-frequency output when the signal carrier is modulated 30 % at 400 cycles per sec.

Would it not be a useful inclusion to the usual standard receiver measurement, as recommended by the Radio Manufacturers' Association, if standardization on " sensitivity via the mains " were added ?

Mr. A. J. Gill and Dr. S. Whitehead (*in reply*):

Degree of Protection.

As regards the degree of protection envisaged, raised by Messrs. Lee, Pawley, and Wynn, the following general considerations should be remembered. The C.I.S.P.R. have only considered so far the correction of non-earthed appliances. B.S.S. No. 800 applies to an item tested in the earthed condition if it can so be used. Accordingly the 500 μV corresponds to a limit of 200 μV or less for non-earthed appliances. Since the chance of an appliance being used in the earthed condition is at present only about 1 in 10 or 1 in 5, since manufacturers must suppress to a lower level in order to avoid rejections, particularly at the low-frequency end, and since a number of appliances are inherently non-interfering, it is clear that only a small proportion will have an interfering voltage as high as 500 μV.

Next there is the joint probability of such a machine being connected at the supply point which has the high coupling factor and of the receiver at the other end being simultaneously operated. It is very difficult to assess the final resultant probability, except to say that it must be very low. In a rough calculation made last year by one of us it was estimated that 96 % of listeners should be protected at present by a 500 μV limit with respect to a field of 1 mV/m., and substantially 100 % with respect to 10 mV/m.; but it is almost certain that the first estimate is pessimistic, and it is safe to say that the number of complaints should, although not eliminated, be very considerably reduced by the adoption of the present proposals.

It appears, from the analysis of certain test areas, that wireless licence-holders and electricity consumers are substantially the same people. To some extent therefore the remedy is in their own hands, and if, as seems likely, B.S.S. No. 800 is voluntarily adopted by most manufacturers, it should require comparatively little publicity to educate the public to require in their own interests the radio-interference-free mark. The Wireless Telegraph Bill now in preparation will no doubt provide some degree of compulsion, but the effectiveness of the measures will ultimately depend to some extent on the co-operation of the public with the manufacturers who are applying correction.

Domestic Appliances.

In connection with power-operated domestic appliances, mentioned by Messrs. Hill, Kinman, Lee, and Struthers, the majority of small motors do not exceed a rating of 500 watts. Vacuum cleaners and polishers are by far the most numerous and, with hair-dryers and fans, cover the great majority. The use of refrigerators is, however, developing, and it is usually found that interference due to the thermostat relay is more difficult to suppress, the motor following known principles.

The effect of the design of the appliance on the interference is complicated, although certain gross defects such as maladjustment of brushes, certain types of bad commutation, and fortuitous connection of parts, are known to increase the interference. A well-made motor should not vary greatly until its brushes are worn to the point of requiring replacement and/or adjustment for efficient running.

A manufacturer can only guarantee correction under the same terms as any other standard of quality, and it is therefore impracticable to envisage other than normal conditions of good usage. Apart from certain simple and avoidable causes of variation, the dispersion of mass-produced articles follows, according to recent tests, a more or less normal law, when the voltage is expressed in db. above 1 μV; and the standard deviation usually lies between 2 and 4. The number having a voltage 5–6 db. above the mean is usually less than 1 %. The high-frequency asymmetry of the machine has a considerable influence on the interfering voltage, and this asymmetry is difficult to reduce for a wide frequency band because it arises from distributed constants. Re-arrangement of windings is sometimes useful, but is not easy to apply to multiple-speed machines. The iron-cored chokes recently developed show a considerable economy in size and weight and should not be a great inconvenience in those instances where the high-frequency impedance of the item is inherently so low as to render them necessary.

We agree with Mr. Colebrook that the term " mains attenuation " is likely to be misleading and we have amended it for the *Journal*.

Methods of Measurement.

We agree with Mr. Wynn that the tolerable signal-noise ratio must ultimately depend on the type of programme, but the listener also enters into the question and, in addition to the C.I.S.P.R. tests, subsequent tests lead to the result that the variation of the mean of a number of observers between different types of programme is less than the variation between different observers for the same type of programme. To obtain a complete solution would involve a lengthy and expensive study, as in the analogous case of telephone interference, but it is difficult to see what method of assessment is possible other than that adopted.

In answer to Mr. Strafford the approximate equality of electric and magnetic vectors has been shown for short-wave and ultra-short-wave interference; on the medium- and long-wave bands, a statistical connection has been made for trolley-buses between the radiation field at a distance and the indication of the standard measuring set at 10 yards. For lifts, attention has been drawn to the fact that the conception both of the interfering and of the broadcast fields inside a building is artificial. The distance of 10 yards was chosen largely for convenience in testing, and the indication is a standard of quality associated with the defined measuring set rather than an absolute measure. The resulting difficulty is mainly associated with the relative liability of receivers to interference, and the open aerial used in measurement is the more common in normal reception.

In reply to Mr. Keall, operation by $f(t)$ gives the instantaneous value of the resultant of the action of an intermittent conductor such as rectification is assumed to be; the load in the detector is assimilated in the filter network constituted by the low-frequency stages. The time-constant of the detector should, in so far as it may modify the audio-frequency response, be corrected by the later low-frequency stages if ideal rectification is to occur. The conditions of operation of the detector cannot differ greatly as between noise and signal, since these conditions are determined in both cases by the carrier, which is assumed to be large compared with either the noise or the modulation. The analysis does not apply to all those departures from ideal rectification, which can, nevertheless, give reasonable reproduction.

Miscellaneous Sources.

We are glad to hear from Mr. Bennett of the success with trolley-buses; trams, being obsolescent, are not always corrected. The error in Fig. 7, noted by Mr. Pawley, has now been corrected.

With reference to neon signs, if Mr. Strafford's objection is to the statement that current passes only during part of a cycle, this feature has been observed in actual oscillograms. While agreeing that relaxation oscillations can be stimulated in the way he describes, oscillograms have also shown, on occasion, damped trains of waves of the type associated with the spark excitation of a multiple resonant circuit.

In reply to Messrs. Forrest and Hill, new designs of insulator show distinct promise, and the types mentioned and others may prove advantageous both as regards normal use and also in avoiding interference. As may be seen from the values quoted, the field from overhead lines is such that interference should only be experienced in close proximity and may be expected to be rare, since such lines are mainly confined to the open country and avoid roads. The current transmitted has little effect on the interference. It is unlikely that interference is transmitted through a high-voltage transformer, although this is possible under some conditions. Distortion in the a.c. supply due to mercury-arc rectifiers is not really a radio problem. It is, nevertheless, very troublesome and difficult to remove. The best measures are to make the capacity of the system large compared with the rectifier load (load mixing), to augment the number of phases on the rectifier, to interpose as many transformers as possible between the rectifiers and the domestic supply, and, finally, to use filter networks, etc., which are often either inefficient or uneconomic.

Short-Wave Interference.

In reply to Mr. Megaw the oscillograms of the current in the automobile ignition discharge, now incorporated in the paper, agree more or less with the maximum observed in the field strength between 40 and 50 Mc., and also confirm his statement that the field strength decreases considerably for still higher frequencies. We appreciate with Mr. Watson the difficulties in the correction of motor cars, but there are a number of possible alternatives, such as to mount the coil on the engine or otherwise to provide a low-impedance path to the chassis, partial screening, resistors, inductors. It should be possible to find an economic solution in a given case, although much development work remains to be done.

Although ignition interference is more widespread, short-wave therapy constitutes the more troublesome problem, but is fortunately limited to certain centres and is intermittently operated. The allocation of certain wavelengths has not been officially proposed, but there is no doubt that a great amelioration of the position could, at least temporarily, be achieved. The medical profession is not, however, decided as to the frequencies which are required, and frequency indicators of sufficient accuracy might prove expensive. Screening can be made effective and portable screens are available abroad, but, even if the latter are satisfactory, treatment of a bed-ridden patient will still prove a difficulty.

The data given by Mr. Megaw are very interesting, and it may be mentioned that quite detailed rules have been drawn up in Canada to deal with electromedical appliances.

In reply to Captain McVicar, expanded metal is more efficient than netting, but the size of mesh cannot be greatly increased owing to the need for the elementary circuits to be of small linear dimensions. For a wide range of frequencies little advantage is to be anticipated from double screening; it is usually more economic to employ the metal in a single continuous screen. Double screening can be disposed for optimum effect, but this depends in a complicated manner on the disposition, nature, and frequency of the radiating source. Window mesh can be bonded by a press joint to the frame and the whole sealed with cement; unfortunately, it is difficult and costly to pass it through a slot and to plumb or sweat. Well-galvanized metal has an indefinite life in dry plaster, but corrosion is likely in damp situations. Corrosion is much accelerated by stray currents, especially in concrete, so that circuits formed by dissimilar metals closed through the surrounding medium should be avoided.

Measures at Listeners' Premises.

In reply to Messrs. Kinman, Hill, Pawley, and Robin, a B.S. Specification for the sensitivity of receivers to interference, the means to reduce this, and the standards to be attained, is now in draft form. Section 3 of the paper gives a few general principles, but it was felt to be premature to deal with all aspects and the issue of the Specification must be awaited. The points raised will, we believe, be adequately treated therein, and the standard envisaged will make a not unreasonable distribution of the burdens as between the parties concerned. The trouble mentioned by Mr. Robin is related to incorrect or defective earthing. The Specification mentioned will contain directions as to earthing, a test of sensitivity to interference at the supply terminals, and a test of the efficiency of filters in reducing this.

Defective wiring is sometimes a cause of trouble, but the chief difficulty, high coupling factors, can occur with sound wiring. It has been decided that it is not economically possible to include the correction of ordinary lighting or power switches.

RESEARCHES IN RADIOTELEPHONY*

By RALPH BOWN, M.E., Ph.D.

(Lecture *delivered before the* Wireless Section, *4th May*, 1938.)

INTRODUCTION

Most of the technical problems which have been encountered in the development of radiotelephony over the past two decades have arisen in one way or another from the necessity of overcoming noise and distortion, two fundamental hazards to telephone transmission. Great research and development effort has been directed toward increasing the power of transmitters and sharpening the selectivity of receivers, simply because these are straightforward ways of reducing noise. In like manner the amplitude and frequency-response characteristics have been studied because perfecting them reduces distortion. In some applications of radiotelephony remarkable quietness and fidelity can now be secured. But in short-wave long-distance service there remain noise and distortion of serious proportions which have their genesis in the vagaries of the transmission medium itself and can best be mitigated by attack upon that part of the system.

One of the earliest weapons used in the attack on the medium was directivity—transmitting directivity to concentrate sending power, and receiving directivity to exclude circumjacent interference. Strangely enough, the newest and most potent weapon brought to bear is again directivity, still sharper directivity, but accurately controlled in accordance with a novel pattern built up by patient research.

DIRECTIVE ANTENNA RESEARCHES

Short-wave directive antennas had not long been in use before it was observed that the improvement they give must be viewed as a statistical average rather than as a fixed value. Occasions were noted when the gain of a directive antenna as compared with a simple doublet was much reduced; in the case of the sharpest receiving antennas the gain might even become negative, as though the antenna were not pointed in the right direction at all. This was taken as evidence of some sort of variability in the angle of approach of the signal, and as a warning that there were practical limits to the sharpening of directivity. Also it was found that the character of the fading and of the accompanying distortion might be different on different antennas; a further indication of idiosyncrasies in the directional behaviour of the signal.

That antenna directivity affected fading and distortion, and that multiple paths of different transit times might be involved, were remarked by C. S. Franklin in 1922 while describing before this Institution some early short-wave directive experiments of the Marconi Co. T. L. Eckersley had already presented proof that ionosphere reflections could explain night errors of radio direction-finding. A few years later Appleton and Barnett, and various other experimenters, settled beyond all doubt the reality of overhead reflections and of multiple, differentially delayed wave components. It also became better recognized that the ground, by reflecting downcoming sky-wave components back up into the antenna, modifies its theoretical " free space " vertical directive pattern, giving it, in the usual case, an upward tilt.

While it may thus be said that more than ten years ago the existence of a relation between antenna directivity, signal gain, and distortion, was generally understood, this understanding was of limited practical value and needed to be implemented by more specific information.

In a series of tests made in 1933 with the kind co-operation of the British Post Office, spurts or pulses of short-wave radiation sent out from Rugby, England, were received at Holmdel, New Jersey, U.S.A., on a simple form of directive antenna system capable of determining the angle, in the vertical plane, between the ground and the incoming ray. By displaying the received signal on a cathode-ray oscilloscope, the number and character of arriving pulses could be seen and recorded. Definite and useful facts were established as follows:—

(1) More than one pulse was usually received and the pulses came in at different times, that is they took different lengths of time to travel from their common source.

(2) The pulses came in at different angles to the earth.

(3) There was a correlation between the times and the angles; the longer the transit time the higher the angle of the trajectory.

(4) The conditions on the individual paths were fairly steady and not subject to fast variations.

These are the conditions which might be expected if the ideal picture of wave paths shown in Fig. 1(*a*) were actually to occur in nature. Fig. 1(*b*) shows how an ordinary simple directive antenna is broad enough in its vertical characteristic to receive several wave components at once. It is well known that the interaction of such wave components produces selective fading and the various forms of distortion resulting from it. Evidently any change in the antenna which modifies its directivity in the vertical plane would affect the nature of the selective fading.

To take advantage of the opportunities offered by this situation, a system known as a " musa " (multiple-unit steerable antenna) has been developed by Messrs. H. T. Friis and C. B. Feldman† and their associates at the Bell Telephone Laboratories. The principles of the experimental musa system in use at Holmdel, New Jersey, are shown schematically in Fig. 2, although many important apparatus details are omitted. Six horizontal

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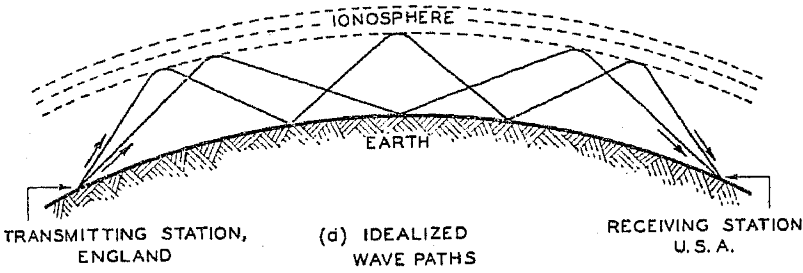
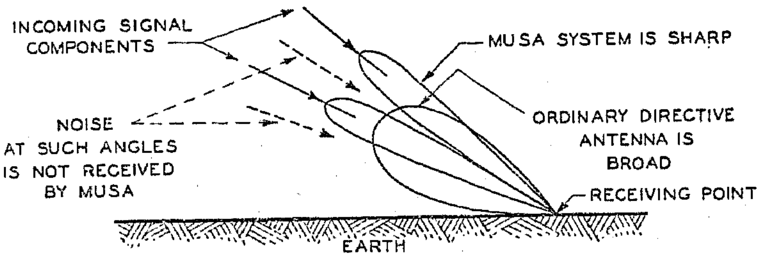
* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 395.

† See Reference (1).



(a) IDEALIZED WAVE PATHS


INCOMING SIGNAL COMPONENTS
MUSA SYSTEM IS SHARP
NOISE AT SUCH ANGLES IS NOT RECEIVED BY MUSA
ORDINARY DIRECTIVE ANTENNA IS BROAD
RECEIVING POINT
EARTH

(b) VERTICAL PLANE DIRECTIVE PATTERNS (NOT DRAWN TO SCALE)

Fig. 1.—Physical basis for "musa" method.

rhombic antennas, arranged in a linear longitudinal array, are connected by coaxial lines to six phase-shifters whose outputs are combined. The phase-shifters are mechanically geared together in such a way that by turning a single control the sharp directivity given by the array may be aimed or steered at any desired vertical angle within the range of the unit antenna characteristics. As is suggested by the diagrams in Fig. 1, the array is so sharp that it may be steered to select one incoming signal ray from another and in this way to receive only signal components of about the same arrival time.

The antennas feed, in parallel, three separate phase-shifter groups, so that three independent outputs, steered for three different receiving angles, may be secured simultaneously. One of the phase-shifter groups is rotated continuously by a motor, to which is also attached a voltage-divider device for furnishing a linear horizontal-sweep voltage to a cathode-ray oscilloscope tube. The detected output of this branch of the receiving system is applied to the vertical deflector plates of the tube. The screen has a slowly decaying phosphorescence and exhibits, for each rotation of the phase-shifter gear, a curve of the relation between signal strength and angle to the earth in rectangular co-ordinates. In this way the angles at which the most effective rays are arriving can be observed, and the other two receiving branches may be set to receive any two desired rays.

The output from each such branch is found to be subject to considerable fading, but it is a much less selective fading than occurs with an ordinary directive antenna. For combining the outputs of the two branches, to gain the advantage of diversity, it has been found sufficient merely to introduce an audio-frequency delay into the branch receiving at the lower angle, to adjust the value of this delay until the two audio signals are in phase, and then to deliver them both to a common output. The delay adjustment is facilitated by a second cathode-ray oscilloscope which has the outputs of the two branches connected, respectively, to its two pairs of plates. Correct delay adjustment produces an inclined line on the screen, while for incorrect adjustment a maze of irregular circles and ellipses is seen.

Two definite improvements are given by musa receiving: (1) The sharp directivity of the receiving branches excludes noise not arriving at the same angles as the received signal components being selected; (2) the separation of the differentially delayed waves and the correction of delay before the components are permitted to combine, reduces selective fading and distortion. In the experimental system at Holmdel 7 to 8 db. improvement in signal-to-noise ratio is realized and the reduction in distortion is frequently quite marked.

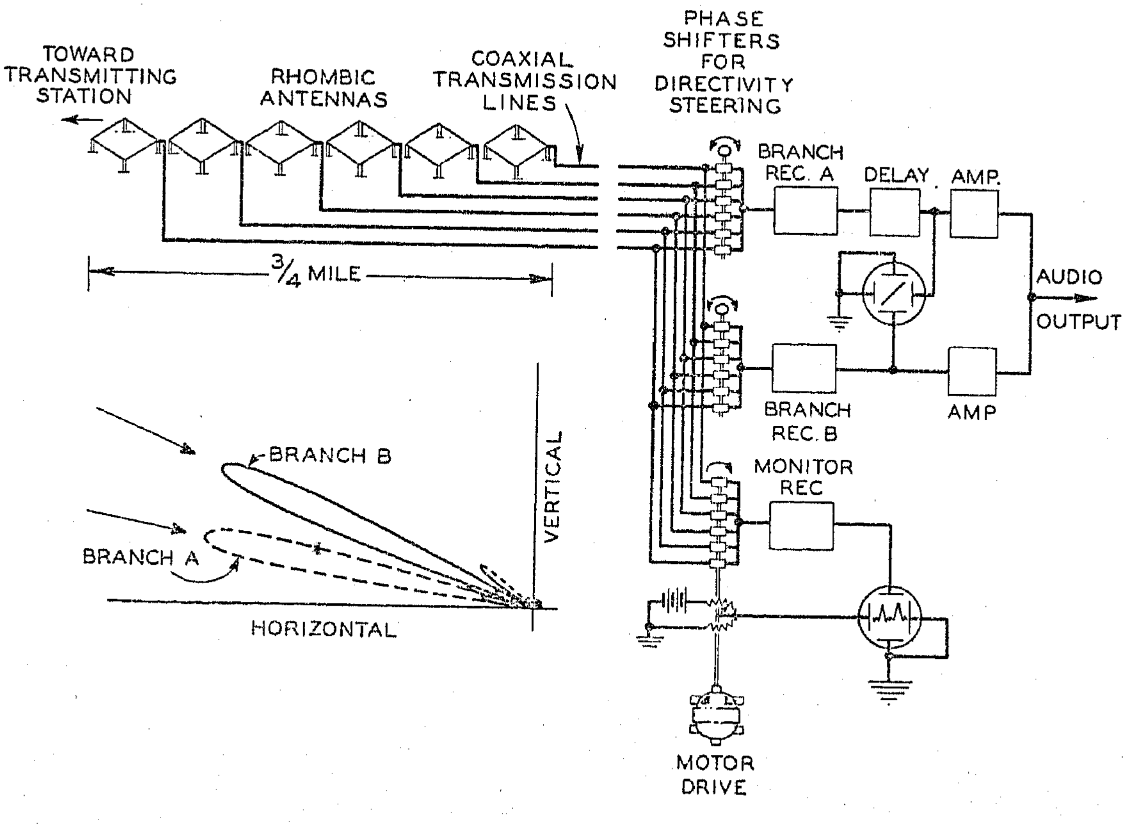


Fig. 2.—Simplified diagram of experimental "musa" system.

CURRENT RESEARCHES IN DIRECTIVITY

With these favourable results from employing sharp steerable directivity at the receiver, it is natural to reason that similar control of directivity at the transmitting end might be beneficial. To explore this possibility a series of transmissions is being sent from Deal, New Jersey, using a simple array of two antennas whose vertical directivity is varied cyclically so as to sweep a null point in the polar radiation diagram repeatedly through a range of vertical angles. Engineers of the British Post Office are co-operating by receiving and recording these signals. Not enough data have been analysed to justify any conclusions but there is found, as one might expect, a suggestion that these east-bound signals follow paths related to the west-bound signal paths which are contemporaneously observed by means of the musa receiver at Holmdel.

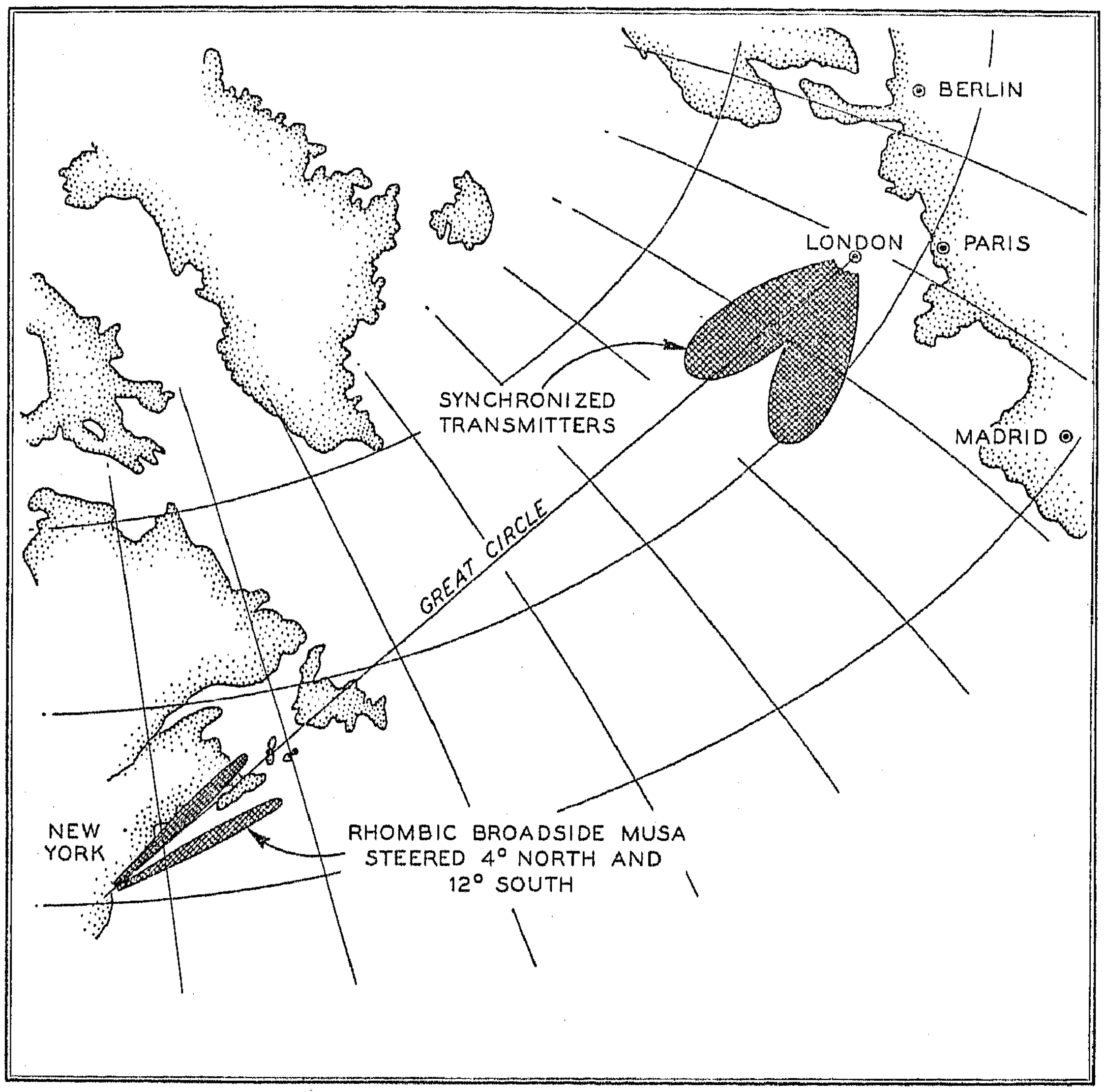

BERLIN
LONDON
PARIS
SYNCHRONIZED TRANSMITTERS
MADRID
GREAT CIRCLE
NEW YORK
RHOMBIC BROADSIDE MUSA STEERED 4° NORTH AND 12° SOUTH

Fig. 3.—Diagram of conditions experienced on the 23rd November, 1937.

While the experiments with the "musa" receiving system have been most gratifying, it has been true that at times the conditions existing seemed to be far from those shown in the ideal diagram of Fig. 1, and the system was unable to cope with them satisfactorily. The evidence, both from the musa tests and from reports of other experimenters, gave ground for suspicion that the difficulty lay, not in unusual behaviour of the wave paths in the vertical plane, but in anomalous performance in the horizontal. At times of severe and very fast fading (flutter fading) observations with simple direction-finding equipment indicated a shower of wave components scattered through a considerable range of angles around the great-circle direction to the transmitter. Following this lead, directional studies are now being pushed into the horizontal dimension, and striking data are being secured with a musa system comprising a broadside array of receiving antennas arranged for sharp steerable directivity in the horizontal plane.

In this work we have not only observed the transatlantic telephone transmissions from the Rugby station of the British Post Office, but also have found a special advantage in the transmissions of the British Broadcasting Corporation from Daventry, because of the variety of transmitting directivity used. Frequently there is simultaneous sending of the same programme in different directions, on different antennas, but on the same frequency. The B.B.C. have been most kind in giving advance information on schedules, antennas, etc. For a comprehensive discussion of this horizontal directivity study I must refer to a recent letter to *Nature* and to a forthcoming paper in the *Proceedings of the Institute of Radio Engineers*, both by Mr. C. B. Feldman.* I would like to mention, however, a few examples of the kinds of conditions which have been observed.

When the path between England and the United States is entirely in daylight, the transmission along the great-circle route is always predominant. This seems to be as true during the abnormal ionosphere conditions associated with solar disturbances as it is in normal times.

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* See Reference (2).

When there is darkness, or partial darkness, on the intervening space, the great-circle route no longer provides the sole transmission path. Frequently there are also waves arriving at other azimuths.

In a considerable number of instances it has appeared as though the ionosphere were warped, there being definite shifts of the direction of wave arrival, sometimes to the north and sometimes to the south, with little regard to the transmitting directivity.

In magnetically disturbed evening periods great-circle paths appear to be attenuated so much that more southern routes often seem to provide the principal means of transmission, even in the case of transmissions directed sharply at New York.

On the afternoon of the 23rd November, 1937, during a moderate ionosphere disturbance, occurred the peculiar circumstances illustrated by Fig. 3. Transmission 4B from Daventry employed two synchronized transmitters, radiating from separate antennas, on 11 750 kc./sec. One was directed at 6° north and the other at 28° south of the true direction to New York. At Holmdel two distinct incoming wave paths were observed, one 4° north and the other 12° south of the great-circle direction to Daventry. Using the broadside musa system these could be separately received. There was a transmission time differential of 1·4 milliseconds between the two paths, the southerly path being longer by this amount.

Much greater deviations have been noticed. For example, at one time during disturbed conditions on the 28th May, 1937, the strongest and steadiest component was 50° to the south of the great circle. A somewhat weaker 5° northerly path also existed.

Although southern paths are usual, they are not always predominant. In the severe magnetic disturbance of 16th–25th January, 1938, marked southern deviations occurred as the magnetic storm developed, and as it receded, but for the period of its greatest intensity a somewhat northerly path was much the most prominent.

Many observations of this sort have been made, and theories to explain them have been formulated, but it is too early to draw practical conclusions. There must be admitted, however, a strong implication that wide-range azimuthal steering of both transmitting and receiving antennas holds promise of improving transatlantic circuits during afternoon and evening hours, particularly when the ionosphere is abnormally disturbed.

SPEECH-CONTROLLED DEVICES

Turning now from directivity to a still older principle for reducing noise interference, I wish again to discuss some recent developments.

The oldest and most direct method of improving signal-to-noise ratio is to boost the signal by increasing transmitter power capacity. When the capacity has been raised to a technical or economic limit, the ingenuity of the engineer is taxed to devise other ways of still further increasing speech power output. One of the most effective of these methods, single-sideband transmission, I shall deal with later. For the moment I wish to consider methods which modify the amplitude characteristics of the speech signal at voice frequencies. Time will permit only a superficial and merely illustrative excursion into this subject.*

Even when there is little inflection in speech, the maximum voltages from a microphone may be as much as 30 db. (30 times amplitude) higher than the significant voltages generated by the weakest sounds. Among different combinations of talkers and connecting lines there may be an additional range of variation of over 40 db. Thus, in commercial systems, a spread of something like 70 db. may actually be encountered between the weakest and the strongest currents it is desired to transmit satisfactorily. The usual way of ironing out the worst of these variations is through manual control of gain by a monitoring operator. The louder conversations or passages of a conversation are depressed, to avoid overloading, and the weaker ones are raised to override the noise. Although it makes a large first-order improvement, this method is obviously incomplete and has all the uncertainties introduced by the human element. The first important improvement to go into commercial radio telephone use was the " compandor," which has operated successfully for some years on the New York–London long-wave circuit.

The compandor does not attempt to substitute for the operator in adjusting the general level of speech; it performs a quite different function which the operator cannot possibly do. The compandor comprises two pieces of apparatus—a compressor at the transmitting end and an expandor at the receiving end of the radio channel. The compressor is a voice amplifier with a fast-acting automatic gain control, which smooths out the speech syllable by syllable and reduces or " compresses " to one half (as measured in decibels) the range of amplitude variation of signals passing through. If the signal thus compressed is used fully to load a radio transmitter, the average volume, and in particular the weakest parts of the signal, will be transmitted at higher level than if no compression had been applied. The weaker sounds in this way override noise more satisfactorily.

At the receiving end the expandor has a sort of *negative* automatic volume control action which doubles (in decibels) the range of variation of signals passing through it. It undoes the work done by the compressor, and expands the speech signal to its original volume range. The net result is an improvement in average signal-to-noise ratio, with no net change in the nature of the signal.

Since the action of the expandor is controlled by the variations of received speech signal itself rather than by a separate pilot channel, spurious changes in intensity caused by fading cannot be distinguished from real signal characteristics and are also exaggerated or " expanded." For this reason the compandor has not been applied to short-wave circuits. A compressor alone is of some value but has not been commercially applied. The expandor, used alone at the receiving end, has a certain beneficial effect, and there has been developed a limited range expandor, sometimes called a " noise reducer," which is better adapted to produce this effect.

The noise-reducer is representative of a class of devices employing a marginal or threshold characteristic. Three different adjustments available in one model of noise-reducer are illustrated by the three curves of Fig. 4. At small values of input, that is for small amounts of noise appearing in the otherwise silent intervals between words or sentences, the gain, and consequently the output, is low. Inputs of higher value cause the gain to rise more or less abruptly to normal value, and full output on actual signals is secured. It is inevitable that some of the weakest parts of the speech will fail to actuate the gain-increasing function and will be lost. If the marginal operating point of the control circuit is correctly adjusted a little above the prevailing noise level, the speech which fails to actuate the control will be so near the noise level as to be of small value in any event.

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* See Reference (3).

While the operation is clearly one of separating the received signals and noise on an amplitude basis, it should be carefully noted that no discrimination is ever achieved between signals and noise which occur simultaneously. The action is simply to suppress both the noise and those speech sounds which are submerged in it. The noise-reducer is at a disadvantage when placed on the output of a circuit subject to fading, because the best margin setting then is not a fixed value but differs from moment to moment; hence a compromise adjustment must be used.

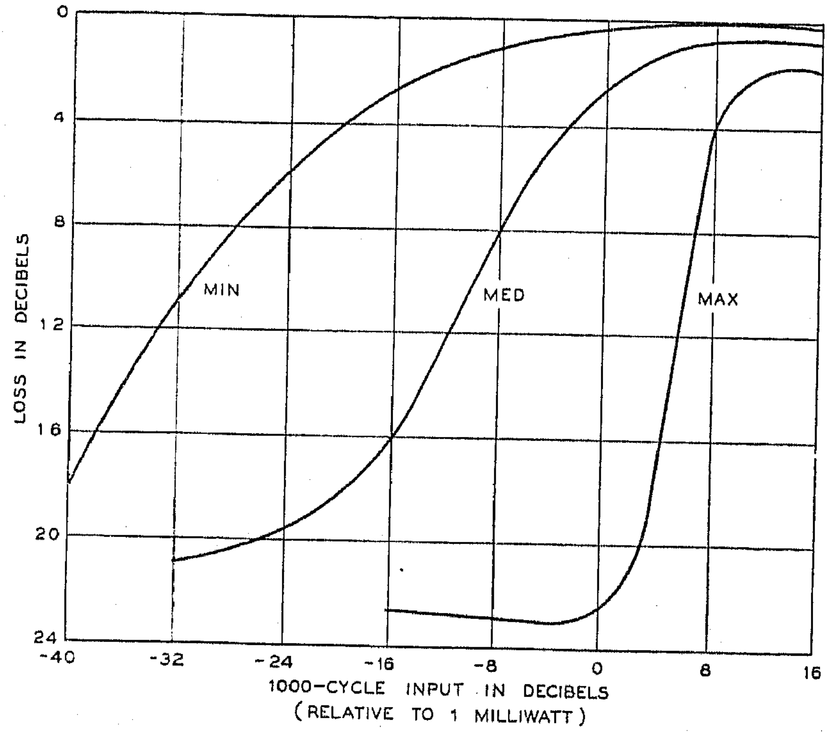

Fig. 4.—Performance curves of a noise-reducer for three different adjustments.

Smoothing out volume variations is a problem upon which important progress is being made. We now have volume-limiters which prevent excessive amplitudes without the distortion caused by the old-fashioned peak-chopper or amplitude-limiter. A volume-limiter has a fixed gain and a linear relation between input and output as long as a selected limiting value of volume is not exceeded. For inputs beyond this limit the gain of the device automatically drops rapidly enough to prevent more than a momentary transient increase of output. The characteristic is again linear at the reduced gain. The gain returns gradually to normal after the excessive input has ceased. The oscillograms in Fig. 5 show how a particular volume-limiter reacts when the input is suddenly increased 10 db. above the limiting value.

In America, volume-limiters of appropriate characteristics are being employed commercially in diverse fields; in radio broadcast transmitter input control, to prevent overloading and permit increasing average programme level; in multiplex wire carrier telephony, to reduce interchannel interference by limiting speech peaks on the individual channel inputs; in short-wave transatlantic single-sideband circuits, to supplement receiver automatic gain control in reducing volume variations.

The volume-limiter curbs excessive upward variations in volume, but as ordinarily used has no effect upon undesirably low volumes. One kind of device which accomplishes both these things has been called a " vogad," a synthetic name derived from the words " voice-operated gain-adjusting device." It attempts to do all that a monitoring operator can do by his listening and by his adjustment of a gain control, and in addition has a certain amount of volume-limiter action. It has two interlocked control circuits. One samples the input speech, and if the volume is low it operates to raise the gain. The other control samples the output speech, and if the volume is up to the desired level it operates to disable the operation of the first control circuit. If the output volume is already above the desired level the second control circuit acts further to decrease the gain. In the case of sudden loud sounds the volume-limiter feature becomes effective until the slower controls can come into action. Except for this volume-limiter function a vogad does not " compress " the speech, that is, it makes substantially no alteration in the moment-to-moment ratios between maximum, minimum, and average voltages of the speech signals. It simply readjusts the gain setting, when needed, to compensate for changes of input volume,* and thus maintain a practically constant output volume. It has the important feature that its gain does not change during pauses in the conversation. The gain remains where it was left by the last words passed, and awaits a recommencement of the talking.

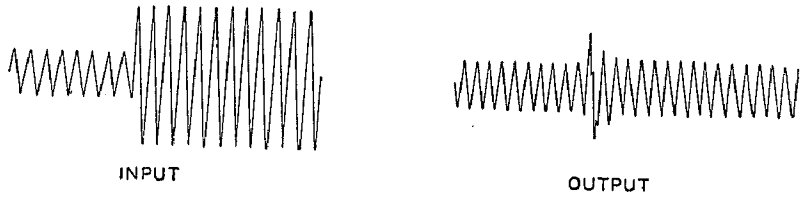

Fig. 5.—Oscillograms showing action of a volume-limiter.

Tests of vogads as against human control on transatlantic circuit operation have shown favourable results. In a ship-shore radiotelephone station under construction at Norfolk, Virginia, a vogad will be employed experimentally to take over the job of regulating volumes into the radio transmitter.

SINGLE-SIDEBAND TRANSMISSION

Single-sideband transmission offers a way of reducing noise and distortion in short-wave telephony which is destined to be of increasing importance.

The single-sideband method, invented in 1915 by John R. Carson, was used in 1917 in wire carrier systems, then considered " high frequency " although the highest band was less than 50 kc./sec. Its application to radio came

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* Volume as defined by the well-known class of visual-reading meter devices called " volume indicators."

10 years later, in the first commercial transoceanic circuit between New York and London. This circuit operates at about 60 kc./sec. and, like the wire systems, employs single sideband with complete suppression of the carrier.

There are two well-known advantages in this method of transmission as compared with ordinary modulated carrier or double-sideband transmission. The first, which is a matter of amplitudes, is that the output of effective signalling power from the transmitting system can be multiplied by four (raised 6 db.) if the carrier is dropped out and all the power of the amplifier is put into the sidebands. The second advantage has to do with band width. By eliminating one sideband and concentrating all the power in the remaining sideband, the occupied band is cut in half and the signalling power per kilocycle is doubled. Assuming that noise is uniformly distributed in frequency and that the receiver is selective enough to exclude all noise except that lying directly in the single sideband intermixed with the signal, an improvement of 3 db. is secured. This makes a total net improvement of 9 db. in signal-to-noise ratio—6 db. at the transmitter and 3 db. at the receiver.

Theoretical analysis of the virtues single-sideband transmission might have where selective fading is present, and experimental testing of its performance on a transatlantic short-wave circuit, were carried out about 10 years ago, but it was not until more recently that complete systems suitable for commercial use were brought forward. In reporting the early experiments it was concluded that single-sideband suppressed-carrier transmission gave less distortion. The conclusion was explained as being due to the fact that a locally supplied carrier was always present in the receiving detector, and it was impossible to have the "blasting" which occurs in double-sideband transmission when the carrier fades out and the sidebands alone remain. Practical operation has now amply demonstrated the reality, not only of the 9-db. gain in signal-to-noise ratio, but also of the improvement in distortion, and the value of the single-sideband system for short-wave telephony seems clearly established.

The first commercial application was to the Netherlands–Java circuit. The Dutch engineers, not content with this achievement, have advanced their work rapidly into the field of multiple channel transmission.

Single-sideband systems have been put into operation between New York and London, and between San Francisco and Honolulu.

There is some divergence of method in handling the carrier frequency. In the Netherlands–Java system, as described in the literature,* the carrier is suppressed and a pilot channel is transmitted at some distance off the upper edge of the speech band, 5 kc. from the carrier position. The system adopted by the American Telephone and Telegraph Co.,† the British Post Office, and others, transmits the carrier in reduced amount, 10 to 20 db. down from normal strength. At the receiver the carrier is separated out and, after being amplified and smoothed or "reconditioned," either is itself fed to the detector or is used to synchronize a local oscillator which in turn supplies the carrier voltage needed in the process of detecting the sideband.

In experimenting with such receiving systems a curious effect has been noted. The carrier is separated from the sideband by passing it through a narrow crystal filter about 40 cycles wide. If, instead of a carrier, a sufficient amount of valve noise or resistance noise is impressed on the input of the filter, its output is still capable of demodulating the speech sideband with little loss of intelligibility, although the detected speech has a peculiar gurgling quality. The resistance-noise impulses have been drawn out by the filter into long transient oscillations of carrier frequency. As might be expected, the gurgling becomes slower if the filter is narrowed still further.

MULTI-CHANNEL PROBLEMS

To secure more efficient use of frequency space in the radio spectrum, it is desirable to progress in the direction of close-packed, grouped channels. In wire and coaxial systems large numbers of telephone channels are sent in a single group, on the scale of 4 kc. per telephone channel. Why can we not do this easily in radio?

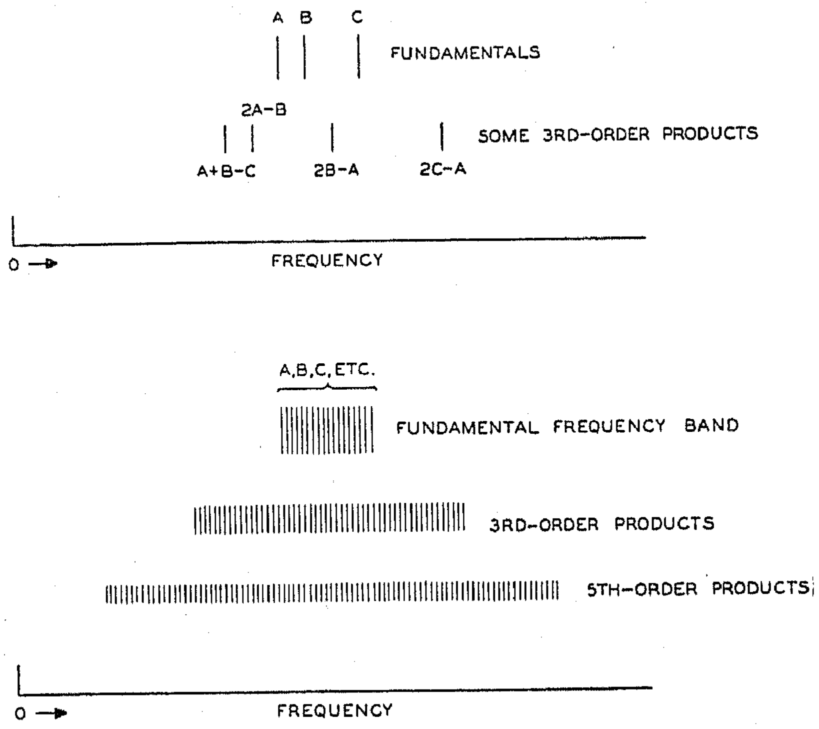

Fig. 6.—Diagram to illustrate odd-order modulation products.

Experiments confirm the expectation that one of the major difficulties is intermodulation between the channels. Theoretical analysis of intermodulation is by no means an easy or straightforward subject. There are, however, a few simple concepts so helpful in visualizing the nature of the problem that their lack of rigour may be overlooked for present purposes.

Objectionable modulation products are caused mainly by distortion in the high-power amplifier stages of the radio transmitter. Such of these products as are well removed in frequency from the signalling bands are of little concern, since they may be adequately suppressed by shielding and by selective output circuits. The so-called odd-order products (3rd, 5th, 7th, etc.) which fall into or close to the fundamental frequency bands, are the seriously disturbing element.

The nature of these products is illustrated in a crude way by Fig. 6. Consider three individual frequencies A, B, and C, representing three components arbitrarily selected from a band of signal frequencies to be transmitted. They are marked "Fundamentals." Intermodulation between them produces new frequencies of the kind A + B − C, A + A − B, 2C − A, etc., in which all combinations and permutations of a sum and difference of any three elements may occur. A few of these are shown in the figure, marked "Some 3rd-order products." Note that they fall both near and among the fundamentals. Now let A, B, and C, be generalized, i.e. moved about so as to represent at one time or another any and all of the frequencies in a signal band to be transmitted. This is illustrated by the lower part of Fig. 6. Evidently intermodulation among a band of fundamentals generates a band of 3rd-order products of three times the width.

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* See Reference (4).

† *Ibid.*, (5).

Three fundamental elements combine to produce one of those spurious frequency elements which we call a 3rd-order product. In a similar way combinations and permutations of 5 fundamental elements can produce spurious frequencies of the 5th order (for example, A + B + C − D − E, or 2A + B − 2C, etc.) covering a band 5 times the width of the fundamental. In general, the higher orders become rapidly weaker. In many cases only the 3rd order is of material importance.

It is interesting to examine the significance of these odd-order modulation products in the problem of conserving radio-frequency space by single-sideband methods.

Consider first the nature of double-sideband transmission as shown in Fig. 7(*a*). Only 3rd-order products are indicated. Spurious frequencies, into the formation of which the powerful carrier frequency enters as a component element, are of greater magnitude than the others. In the diagram 3rd-order products which can be of this kind are indicated by longer vertical lines.*

If one sideband is eliminated and the carrier is reduced and the remaining sideband is increased, the condition changes to that shown in Fig. 7(*b*). The band occupied by the fundamental frequencies becomes narrower in the ratio of 2 : 1, and that occupied by the 3rd-order products also contracts 2 : 1.

Experimental measurement of distortion products on an actual transmitter checks fairly well with this simple analysis, as may be seen from Fig. 7(*c*). Here are plotted the distortion products picked up by a sharply selective single-sideband receiver when tuned with its mid-band frequency displaced from the carrier position by different amounts. One curve is for ordinary double-sideband transmission; the other is for single-sideband reduced carrier transmission. In both cases the modulating speech was inverted. The same peak load on the high power amplifier was used for both kinds of transmission. The ordinates are in decibels referred to the single-side band signal correctly tuned in. By their wide spread these curves indicate that 5th-order, and possibly even higher-order, products are present in measurable magnitude.

It is customary to establish the adjustment and loading of transmitters to meet limits of signal distortion which do not curtail intelligibility and do not offend the ear. Study of experimental evidence shows that something more than this is necessary to permit packing radio channels as closely as channels are packed in wire circuits. The modulation products which fall back into the band occupied by the fundamental frequencies become inextricably mixed with the signal and are recognized at the final output as signal distortion. The fringe products, which fall outside the fundamental band, may spread into neighbouring channels of communication if these lie close enough. Products falling back into their originating channel as signal distortion may be tolerated up to a magnitude of perhaps 5 % of the fundamental (25 db. down); for products which appear as cross-talk in another telephone channel a value of less than 1 % of the signal in that channel is desirable (more than 40 db. down).

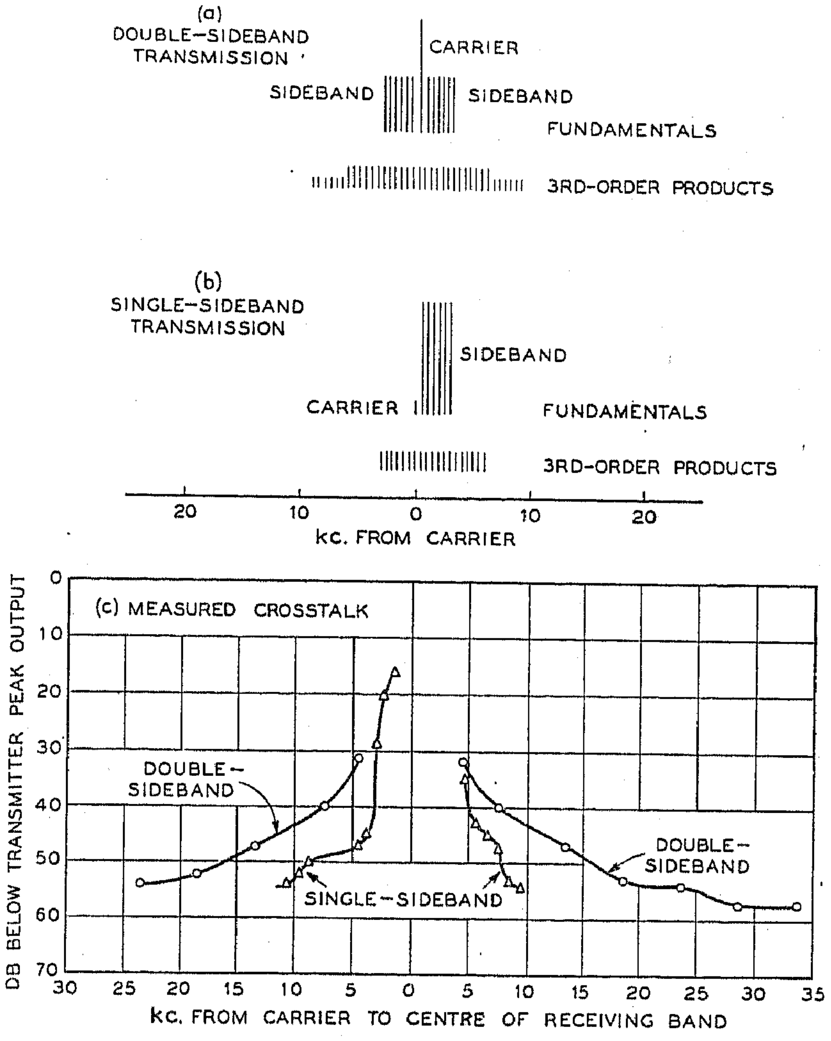

Fig. 7.—Comparison of double-sideband and single-sideband modulation products.

Assuming a tolerable cross-talk ratio of 40 to 45 db., it might appear, from the single-sideband curve of Fig. 7(*c*), that channels of equal powers could be spaced 5 or 6 kc. between centres. The conclusion is justified only if it is certain that the modulation products from one channel (channel "x") which fall into any other channel (channel "y") have fading variations equal to and contemporaneous with those experienced by the fundamental components of channel "y." This condition can be produced with certainty only by radiating channel "y" and the modulation products of channel "x" both from the same antenna so that they will follow identical transmission paths. If they are radiated from separate antennas, or from different stations, channel "y" may fade down when the modulation products from channel " x " fade up, and the cross-talk ratio may, for a considerable part of the time, be intolerable.

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* The diagrams (*a*) and (*b*) illustrate band widths primarily and are not intended to be more than crudely suggestive of relative amplitudes.

The method at present most available for introducing several closely spaced channels at higher power into a common antenna is to aggregate the channels into a group at low power and then to amplify them in a common power amplifier which feeds the antenna. In the amplifier not only does each channel generate its own modulation products as pointed out above, but additional products are generated by intermodulation between the different channels, and the cross-talk and interference problems become much more complicated.

In wire systems handling many channels in a group, the amplifier valves can be operated with large amounts of negative feedback to reduce intermodulation to very small values. Whether it will be possible to obtain satisfactory efficiency and freedom from distortion in high-power radio-frequency amplifiers will depend not only upon problems imposed by the greater magnitude of power and frequency, but also upon the adaptability of the necessary circuits to practical operating requirements, such as quick and easy wave-changing.

Presumably the goal will not be achieved in one stroke, but will be approached by various practical compromise schemes which make good use of the available art. The 2- and 3-channel arrangements tried out on the Netherlands–Java circuit, and the twin-channel system being tested between New York and London, are examples of progress along this line. At any rate we may be sure that research workers will continue to do their part toward making it more difficult to be pessimistic about the limitations of radio-frequency space.

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(This paper, which is a comprehensive survey of the subject with an extended bibliography, was presented at the Pacific Coast Convention of the American Institute of Electrical Engineers at Portland, Oregon, 11th August, 1938, and is expected to be published in due course in the *Bell System Technical Journal*.)

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A METHOD OF USING HORIZONTALLY POLARIZED WAVES FOR THE CALIBRATION OF SHORT-WAVE FIELD-STRENGTH MEASURING SETS BY RADIATION*

By J. S. McPETRIE, B.Sc., Ph.D., Associate Member, and B. G. PRESSEY, B.Sc.

[From the National Physical Laboratory.]

(*Paper first received* 17*th November*, 1937, *and in final form* 22*nd February*, 1938.)

SUMMARY

The paper describes a method for the calibration of short-wave field-strength measuring sets by radiation using a loop transmitter in the horizontal plane. It is shown experimentally that in the case of vertically polarized waves† the simple ray theory does not apply unless the transmitter and receiver are both elevated to considerable heights above the ground. With horizontally polarized radiation, however, the simple ray theory holds on short waves for practically all heights of transmitter and receiver. This distinction between the propagation characteristics of the two types of radiation suggests the use of horizontally polarized waves for field-strength calibrations on short waves. The advantages ensuing from the use of horizontally polarized waves for this purpose may be summarized as follows:—

(*a*) The heights of transmitter and receiver may be reduced practically to any extent, so that high masts are no longer required.

(*b*) For low heights of transmitter and receiver the reflection coefficient of the ground is sensibly unity for all types of the earth's surface, so that little error can be introduced in the analysis due to wrong assumptions as to the electrical properties of the ground.

The analysis of the results is considerably simplified when the reflection coefficient of the ground is approximately unity.

(*d*) With no masts at the transmitter or receiver it is an easy matter to take an attenuation run, so that the calibration is not dependent on one observation.

(*e*) It is particularly suitable for the calibration of receivers incorporating rectilinear antennae, as there is no error in calibration due to varying height of the aerial above ground.

(1) INTRODUCTION

The enormous increase in the congestion of the radio channels on long and medium wave-bands has forced the radio engineer to investigate the possible uses for communication of the ultra-short waves less than 10 metres in wavelength. In order to determine the commercial possibilities of such waves a knowledge of their propagation characteristics is essential and this, in turn, entails a method of measuring field strength. The measurement of field strength at any frequency depends ultimately on the measurement of voltage or of the current through a known impedance. Even when voltage and current can be measured at high frequencies, numerous difficulties are encountered in practice, owing, amongst other things, to the disturbance introduced by the measuring instrument itself. Until an accurate and convenient method of measuring current is available for short waves, recourse must be had to the thermo-junction, the calibration of which with low-frequency alternating current is assumed to hold at the highest frequencies. It has been shown† recently that, with existing types of thermo-junction, the probable error introduced in this way in the 5- to 10-metre region is not more than 5 %. This error is sufficiently small for present-day purposes on this waveband.

A small vertical loop transmitter has been used by American investigators‡ to produce by radiation an electromagnetic field in the neighbourhood of the receiver to be calibrated. By measuring the current in the transmitting loop of known area the field at the receiver can be calculated, allowing for the effect of the reflection at the ground. This ingenious method has the great advantage over the methods usually employed on longer wavelengths when applied to the ultra-short wave-band that no extraneous leads or impedances of doubtful magnitude are added to the receiver during calibration. In order to minimize the effect of the ground, and so the error due to possible wrong assumptions as to its reflecting properties, the transmitter and receiver are usually placed approximately one wavelength above the ground and about half a wavelength apart. This entails the use of relatively high masts, and the calibration is dependent on one disposition of transmitter and receiver. With the radiation method described below, using horizontally polarized waves, the receiver calibration can be made for low heights of transmitter and receiver. The subsequent analysis of the results is considerably simplified, and by measuring the attenuation of field strength with distance the calibration can be made dependent on a series of observations. It is shown in the paper that the calibration obtained in this way is the same as that with vertically polarized radiation between transmitter and receiver, each of which is at a height above the ground comparable with the wavelength.

(2) DESCRIPTION OF TRANSMITTER

A loop transmitter was used because of the uniform radiation in all directions in the plane of the loop, when the perimeter of the latter is small compared with the wavelength. The current in the loop which constituted

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* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 210.

† Throughout this paper, the expressions "vertically" and "horizontally" polarized waves are used to mean radiation of which the electric vector lies in and perpendicular to the plane of incidence, respectively.

† See Reference (1).

‡ *Ibid.*, (2).

the inductance of the oscillatory circuit of the transmitter as well as the radiating source was measured by a non-contact type of thermo-junction the couple of which was connected by a twisted pair of No. 40 gauge copper wires to a microammeter of small dimensions. The radiation from such a transmitter is similar to that of a magnetic doublet at the centre of the loop. If the loop is placed near any reflecting surface, such as a metal box containing the radio components or battery supply of the oscillator, an image of the primary radiating source is formed in the box. The effect of this image is to alter the radiation distribution of the transmitter to an indeterminate extent. For this reason all screening of the oscillator was dispensed with and the radiation due to high-frequency currents in parts of the oscillator other than the loop was reduced to a minimum by making the physical size of all components of the oscillator, except the loop, as small as possible. In order to maintain electrical symmetry in the loop so that the "antenna effect" of the loop and the high-frequency current in the high-tension lead should both be small, the push-pull circuit as shown in Fig. 1 was used, incorporating two acorn valves. The battery leads were arranged perpendicular to the plane of the loop, and the batteries (of an unspillable type) were lashed to a platform parallel to the plane of the loop and distant from it by about 4 ft. By this disposition of the components of the transmitter the polarization of the radiation from high-frequency currents in the battery leads was in a plane perpendicular to that of the loop and could not, therefore, induce any e.m.f. in the receiver loop when the latter was orientated so as to receive the radiation from the transmitter loop.

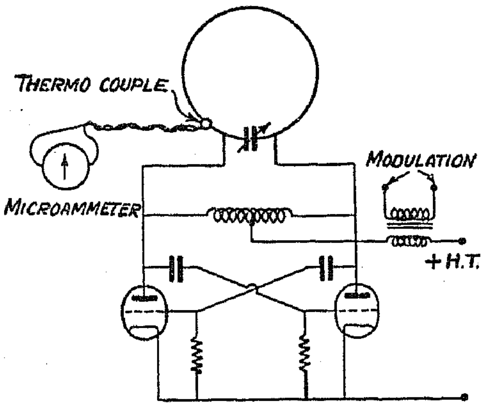

Fig. 1.—Circuit diagram of transmitter.

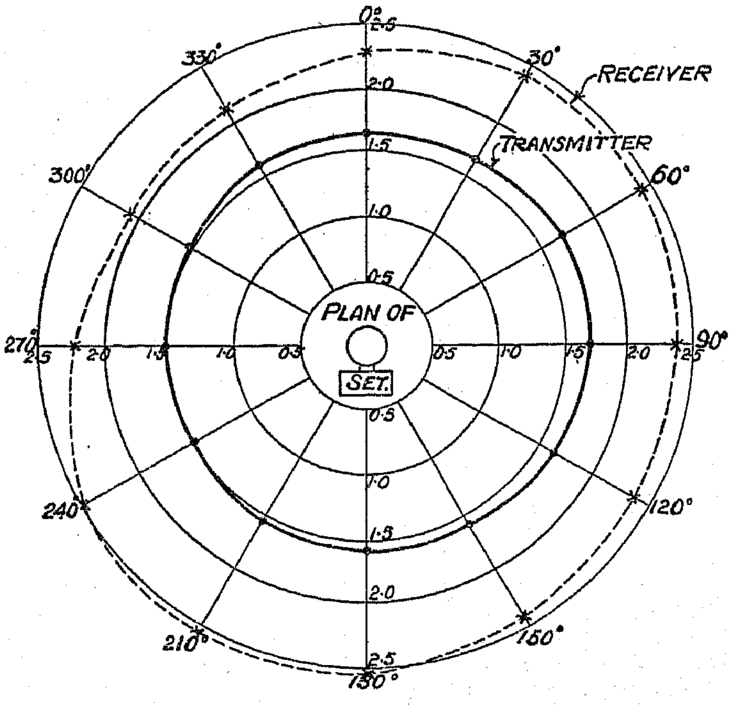

Fig. 2.—Polar diagrams of transmitter and receiver.

The usual law for the radiation from a loop source assumes that the current throughout the loop is uniform. The degree to which this condition holds in any particular case can be determined by finding the magnitude of the departure from uniform radiation in all directions in the plane of the loop. On account of the reflection occurring at the earth's surface the most convenient way to make this test is to place the radiating loop horizontally and to measure the horizontal field at a fixed receiver for different orientations of the transmitting loop. This test was made on all wavelengths for which the transmitter described above was used as a standard radiating source. For a given loop perimeter it was found, as would be expected, that the variation with direction of the received signal from its mean value increased with decrease in wavelength, but with the loop sizes used the total variation was always less than 10 % of the mean field. It can be assumed, therefore, that the calculation of field from the usual formulae involving the loop dimensions and current, hold to at least the same accuracy. The full-line curve in Fig. 2 shows the polar radiation distribution found in this way for a loop perimeter of 0·96 m. on a wavelength of 8 m. (35·5 Mc./sec.).

(3) DESCRIPTION OF RECEIVER

The receiver was of the superheterodyne type incorporating a single-turn loop as the aerial and also as the inductance of the first circuit. A block diagram of the receiver is given in Fig. 3 so that the disposition and dimensions of its components may be appreciated. The small metal box directly under the aerial loop contains the frequency-change unit consisting of two diodes in push-pull, with appropriate beat oscillator diodes being used so that overloading in this stage of the receiver was prevented. The amplitude of the local oscillator voltage was sufficient to ensure that the intermediate frequency output was independent of this voltage. The output from the frequency-change stage is transferred by means of a screened lead to the intermediate-frequency amplifier contained in the large metal box. The wooden box at the bottom of the receiver contains all the batteries. Before making a measurement of field strength the gain of the intermediate-frequency amplifier was adjusted by varying the grid-bias of the valves, so that with the frequency-change unit disconnected the thermal-agitation noise in the first intermediate-frequency circuit produced a standard reading of the output meter. The frequency-change unit was then connected to the amplifier, tuned to the frequency of the radiation on which a measurement was required, and resistance attenuators in the intermediate frequency amplifier were adjusted so that the reading of the output meter was the same as that given by first-circuit noise and no attenuation. By the initial adjustment of the intermediate-frequency amplifier, the overall amplification of the receiver could always be set to the same value, since the use of diodes in the frequency-change unit allowed of no alteration in the response of that part of the receiver.

A loop aerial was used in the receiver because in the experiments to be described two waves from the transmitter were incident at the receiver, one by the direct path and the other after reflection at the ground. For elevated transmitter and receiver these two waves were not in the same azimuth, so computation could be reduced considerably by using at the receiver a loop aerial having uniform " pick-up factor " for all directions in its own plane. In order to test whether the large metal boxes of the receiver affected the circular polar diagram of the receiving loop, the receiver was placed with its loop in the horizontal plane and all its components arranged with respect to each other and the plane of polarization of the incident radiation in exactly the same manner as for the reception of vertically polarized waves. The height of the receiving loop above the ground was 1·2 m. and the receiver was so arranged that it could be tuned by means of strings by an operator at a distance of about 2 m.

The horizontal field at the receiver was measured as the transmitter with its loop arranged horizontally at a height of 0·84 m. was moved to a series of positions 30° apart at a constant distance of 25 metres from the receiving loop. In order that the polar radiation distribution of the transmitter should not affect the results, the transmitter loop remained in a fixed direction at each position relative to the radial line to the receiver. The observations were repeated with the receiver rotated through 180° about the vertical axis through its loop. By taking the average of the two measurements (the maximum difference was about 10 %) for the same position of the transmitter relative to the receiver loop, the effect of any spurious reflections from nearby conductors was reduced. The results of measurements made at a wavelength of 8 m. are shown by the dotted curve in Fig. 2, from which it is seen that the maximum variation of received signal from its mean value is ± 7·5 %. This variation was sufficiently small for the radiation reception distribution of the receiving loop to be considered uniform.

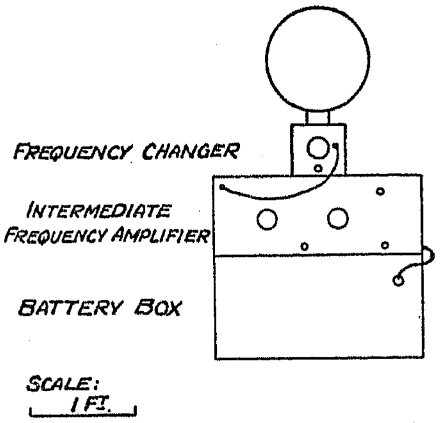

Fig. 3.—Scale diagram of receiver.

(4) SIMPLE RAY THEORY OF WAVE PROPAGATION

The radiation in the immediate neighbourhood of a transmitting aerial is extremely complicated. At distances from the source greater than about one wavelength, however, the radiation may be considered as that of a plane wave. The earth behaves in many ways as a radiating source when radiation is incident on its surface. For this reason the plane-wave theory may not apply if the radiation is required from a transmitter placed near the earth at a receiver also at a low height. Let T, Fig. 4, represent a source of strength I at such a height h_1 above the ground that the radiation incident on the earth may be considered as plane, and R that of a receiver also at an appreciable height h_2. The radiation incident at R can be considered as that due to the primary source I at T and a secondary source at the image point of T in the earth. The strength of this secondary source is $\rho f(\theta) I$, in which ρ is the reflection coefficient of the ground for plane waves of the appropriate frequency and angle of incidence θ, and $f(\theta)$ represents the polar radiation distribution of the source at T. The magnitude of the reflection coefficient and the phase change on reflection for any condition of reflecting surface and angle of incidence can be determined from a series of curves given by one of the authors.* If the source at T is a vertical single-turn loop the value of $f(\theta)$ is unity and the field at the receiver is then given by

$$E = \frac{120\pi^2 AI}{\lambda^2}\left[\frac{1}{a} + \frac{\rho}{b} e^{\frac{2\pi j}{\lambda}(b-a)}\right] \quad . \quad . \quad (1)$$

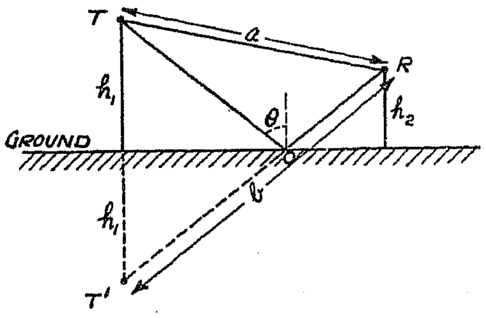

Fig. 4.—Diagram showing (*a*) direct and (*b*) reflected wave equivalent paths between elevated transmitter (T) and receiver (R).

in which A is the area of the radiating loop, I the current in the loop in amperes, λ the wavelength, and a and b the distances TR and T′R. The field is given in volts per unit of the dimension of length used in equation (1). When the product $h_1 h_2$ is very small compared with the horizontal distance D between transmitter and receiver

$$\rho \rightarrow -1$$

$$\frac{1}{a} \simeq \frac{1}{b} \simeq \frac{1}{D}$$

and $$(b-a) \simeq \frac{2h_1 h_2}{D}$$

Equation (1) then reduces to

$$E = \frac{480\pi^3 AIh_1h_2}{\lambda^3 D^2} \quad . \quad . \quad . \quad . \quad (2)$$

Equation (2) should give the attenuation of vertically polarized waves along the ground if the conditions for plane-wave propagation apply. It can be shown that equation (2) is equally true for horizontally polarized waves under the same conditions.

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* See Reference (3).

(5) EXPERIMENTAL PROCEDURE AND RESULTS

(a) Vertically Polarized Waves

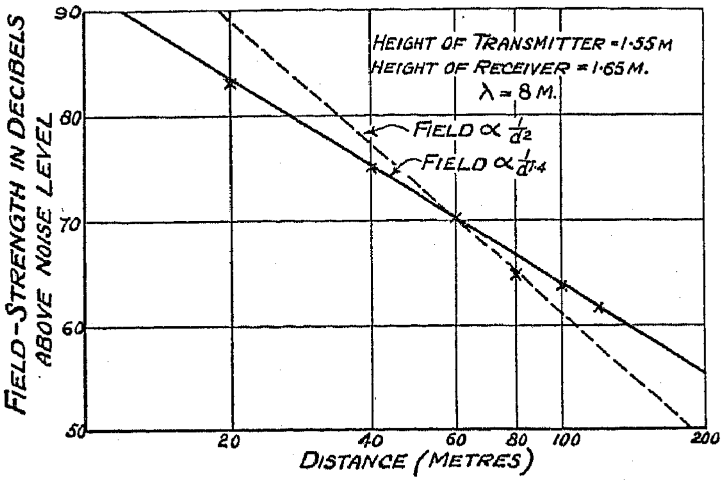

Fig. 5.—Variation of field strength with distance for vertically polarized waves.

The propagation of vertically polarized waves has been the subject of extensive study since the classical work of Zenneck[*] and Sommerfeld.[†] While much experimental work has been carried out previously on the attenuation of waves along the ground for longer wavelengths, it was considered desirable, for the purpose of this paper, to show experimentally that the attenuation of vertically polarized ultra-short waves at low angles of elevation along the ground is not given by equation (2), from which the field from a given transmitter should vary inversely as the square of the distance between transmitter and receiver, provided the product of transmitter and receiver heights is small compared with that distance. In order to test whether this condition applied to the propagation of vertically polarized waves, the receiver described in Section (3) was set up with its loop in the vertical plane and at a height of 1·65 m. above the ground. The transmitter (Section 2) with its loop in the same vertical plane as that of the receiver loop, and at a height of 1·55 m., was placed in turn at various distances between 20 and 120 m. from the receiver. The field strength at the receiver in relative values was determined for a constant current in the transmitting loop operating on a wavelength of 8 m. (37·5 Mc./sec.). The field strength measured in this way is shown on logarithmic scale in Fig. 5 plotted against the corresponding distance. The dotted line in the same figure is the inverse-square law given by equation (2). The departure from the experimental values of field strength given by this relation is greater than that which can be accounted for by experimental error. The full line, which is the best mean through the experimental points, satisfies the relation[‡]

$$\text{Field} = \text{constant}/(\text{distance})^{1\cdot 4} \quad . \quad . \quad . \quad (3)$$

According to equation (2) the field strength should be proportional to the height of either transmitter or receiver. Measurements made with a distance of 60 m. between transmitter and receiver showed, however, that no change of field strength occurred when the height of the transmitter was increased from 1·5 to 2·25 m., while the receiver height remained fixed at 1·65 m.

The results obtained show, therefore, that equation (2) cannot be used to determine the field at low elevations for vertically polarized waves. It was pointed out above that if the transmitter and receiver are both elevated to considerable heights from the ground the re-radiation from the earth is plane at the receiver and the laws given by the ray theory are correct. In order to calibrate the receiver with vertically polarized waves, the receiver was located at the top of a 20-ft. wooden-lattice tower. The transmitter was raised to a height of 40 ft. on a light wooden mast which could be easily moved from one position to another. The transmitter loop was supported on a 3-ft. pole projecting horizontally from the top of the mast in a direction perpendicular to the plane of propagation. Relative measurements of the field at the receiver were made for horizontal distances between transmitter and receiver of from 10 to 40 m.

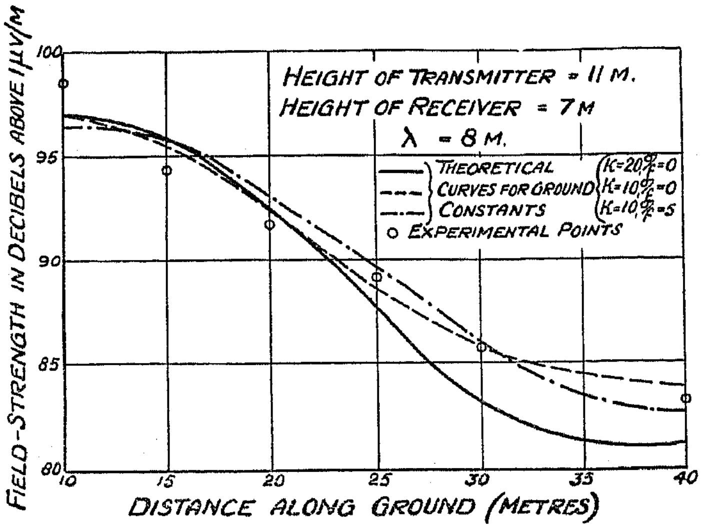

Fig. 6.—Variation of field strength with distance for elevated transmitter and receiver for vertically polarized waves.

The calculation of the field strength at the receiver necessitated the use of equation (1), as the heights of the transmitter and receiver were no longer small compared with the distance between them. The only way in which the magnitude of the reflection coefficient in this expression could be determined was to assume values for the dielectric constant κ and the conductivity σ of the earth's surface at the frequency f of the radiation used, determine the value of ρ from Fresnel's equations for the angles of incidence corresponding to each position of transmitter, and find which values of these constants gave the curve in Fig. 6 nearest in shape to that obtained experimentally. The values of earth constants at 8 m. considered as limiting values and taken to determine ρ were $\kappa = 20$, $\sigma/f = 0$; $\kappa = 10$, $\sigma/f = 5$; and $\kappa = 10$, $\sigma/f = 0$.

The calibration of the receiver required was to determine the field which would give the standard output with no signal attenuation. Assuming this field to be 1·8 μV/m., the experimental results obtained are shown by the circles in Fig. 6. It will be seen that they lie closely to the theoretical curves for $\kappa = 10$ and $\sigma/f = 0$ or 5.

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[*] See Reference (4).

[†] *Ibid.*, (5).

[‡] This relation is approximately the same as that found by Burrows for short-distance propagation over water. He states that this variation of received field strength agrees with that derived from the formulae given by Weyl and Norton (C. R. BURROWS: "The Surface Wave in Radio Propagation over Plane Earth," *Proceedings of the Institution of Radio Engineers*, 1937, vol. 25, p. 219). The electric field for vertically polarized waves is not given accurately in magnitude by equation (2), but over the range of distances used in these tests the variation should, according to the ray theory, be approximately inversely proportional to $(\text{distance})^2$.

Therefore, the magnitude of a field which produces the standard output with n decibels attenuation in the receiver is n decibels above $1\cdot 8\ \mu V/m$.

(b) Horizontally Polarized Waves

The departure of the law of radiation observed with vertically polarized waves at large angles of incidence from that given by equation (2) is to a large extent due to the rapid variation with angle of incidence in the reflection coefficient of the ground for such waves. The full and dotted curves in Fig. 7 give the reflection coefficients for ground constants $\kappa = 10$, $\sigma/f = 0$, for vertically and horizontally polarized waves respectively, for angles of incidence between 86° and 90°. It will be seen that the rate of change of reflection coefficient with angle of incidence is much less for horizontal than for vertical polarization. This result suggests that equation (2) might apply for horizontally polarized radiation for relatively small heights of transmitter and receiver. In order to test this the receiver was set up with its loop horizontal as in the experiments described in Section (3).

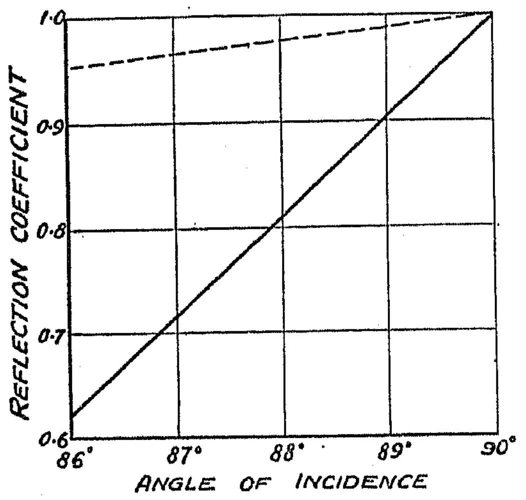

Fig. 7.—Reflector coefficient/Angle of incidence for ground constants $\kappa = 10$, $\sigma = 0$.

- - - - - - Horizontal polarization.
———— Vertical polarization.

The transmitter with its loop also horizontal was placed with respect to the receiver so that its radiation was incident on the receiving loop in the same azimuth as vertically polarized radiation from a distant source when the receiver was used for the reception of such waves. The heights of transmitter and receiver were respectively 1·20 and 1·15 m. The field strength at the receiver was measured relative to noise-level with the transmitter at various distances between 20 and 140 m. The experimental results obtained on a wavelength of 8 m. are plotted in Fig. 8 as field strength in decibels above noise-level in the receiver against the distance on a logarithmic scale. The points lie within experimental error on the full line in the same figure, which is given by equation (2), viz.

$$\text{Field strength} \propto \frac{1}{(\text{distance})^2}$$

Equation (2) also shows that if the ray theory can be applied at low angles of elevation the received field should be proportional to transmitter height, other factors remaining constant. The transmitter and receiver with loops horizontal were placed 100 m. apart, and while the receiver height remained fixed at 0·8 m. the transmitter height was varied in steps from 0·6 to 2·7 m.

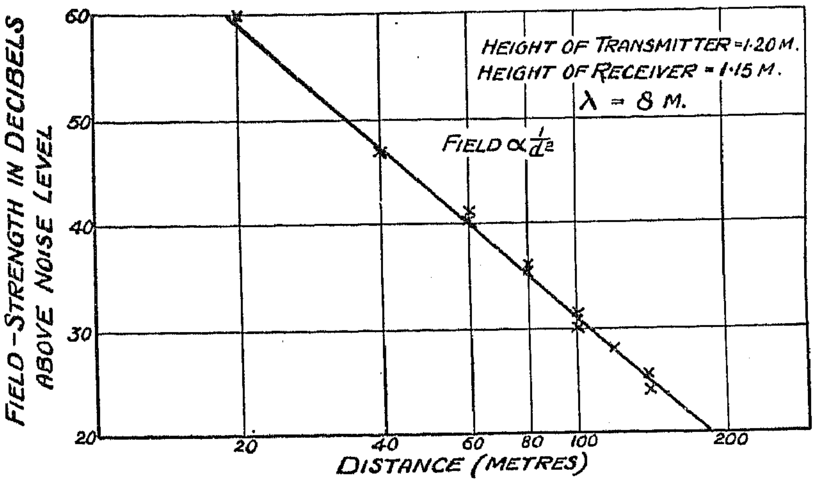

Fig. 8.—Variation of field strength with distance for horizontally polarized waves.

For each height of transmitter the field strength at the receiver was measured for a constant current in the transmitting loop. The field strength in decibels above noise-level is plotted in Fig. 9 against the logarithm of transmitter height. The full line is given by the relation

$$(\text{Field strength}) \propto (\text{Transmitter height})$$

The experimental points lie closely to this line, so that the field strength at the receiver is proportional to transmitter height for horizontally polarized waves. It follows from the theorem of reciprocity that a similar result would have been obtained if the height of the receiver instead of the transmitter had been changed. This means that the field strength at a low receiver from a low horizontal transmitter is proportional to the product of transmitter and receiver heights.

The measurements with horizontally polarized waves show, therefore, that the propagation of such waves is in agreement with equation (2), which was arrived at on the assumption of plane-wave propagation. The logical deduction from this result is that, to the accuracy obtainable with the receiver used for the measurements, the conditions of plane-wave propagation can be assumed with horizontally polarized waves for all practical heights

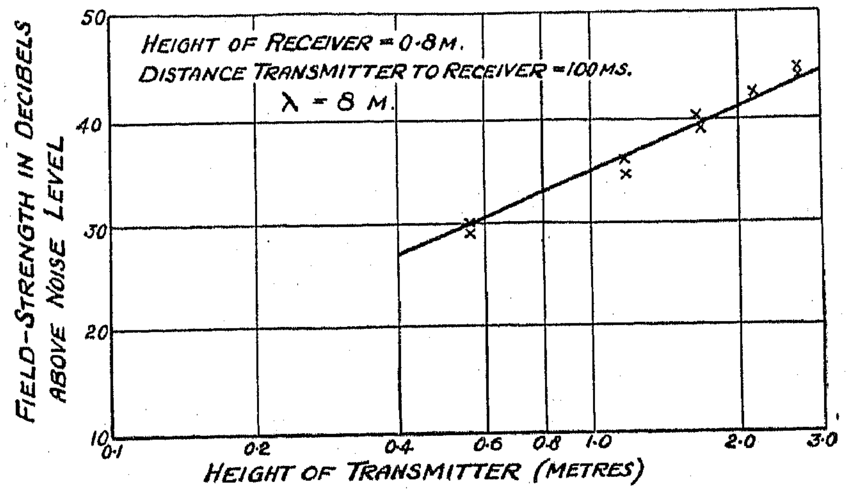

Fig. 9.—Variation of field strength with height of transmitter for horizontally polarized waves.

of transmitter and receiver. The field at a distance of 100 m. for the transmitting conditions obtaining in the experiment calculated according to equation (2) is 64·3 μV/m. The corresponding reading of the field strength measured at the receiver in decibels above noise-level is 31·5. Hence with the horizontal method of calibration by radiation the field strength corresponding to zero receiver attenuation for the standard output is 31·5 db. below 64·3 μV/m., that is 1·7 μV/m.*

(6) COMPARISON OF THE TWO METHODS OF RECEIVER CALIBRATION

The closeness with which the experimental results given in Fig. 6 can be made to agree with the calculated values of field strength shows that under the conditions specified the assumption of plane-wave reflection at the ground was justified. The calibration using elevated transmitter and receiver should, therefore, be correct within experimental error. The field strength equivalent to the standard output determined in this way was 1·8 μV/m.

With the method using horizontally polarized waves, the results given in Figs. 6 and 7 show that the relation between field strength and transmitter and receiver heights and distance is that obtained in equation (2). The magnitude of the field under given conditions need not, however, be given by this equation. The calibration of the field strength for standard output, assuming equation (2) to be correct, was 1·7 μV/m., which agrees within experimental error with the value 1·8 μV/m. obtained with vertically polarized waves. The horizontal method of calibration with small transmitter and receiver heights and the supposition of plane-wave reflection may be used, therefore, for the calibration of short-wave receivers.

(7) CONCLUSIONS

The investigation described above leads to the following conclusions.

(*a*) The simple ray theory does not apply to the propagation of vertically polarized waves directly along the ground.

(*b*) In order that the simple ray theory may be applied for short distances the transmitter and receiver must be raised off the ground to heights comparable with the wavelength.

(*c*) The computation of the field under the conditions specified in (*b*) is laborious, and the accuracy of the result is dependent on correct assumptions as to the electrical properties of the ground.

(*d*) The simple ray theory applies to the propagation of horizontally polarized waves of wavelengths less than 10 m. for all practical heights of transmitter and receiver.

(*e*) Result (*d*) suggests the use of horizontally polarized waves for the calibration of short-wave receivers by radiation. Low transmitter and receiver heights can then be used. The computation of the field at the receiver under this condition is very easy and accurate, as at grazing incidence the reflection coefficient is practically unity for all types of ground.

(*f*) The calibrations of a receiver using vertically and horizontally polarized radiation for high and low transmitter and receiver heights, respectively, agreed within 10 %.

(8) ACKNOWLEDGMENTS

The work described in this paper was carried out as part of the programme of the Radio Research Board, and acknowledgment is due to the Department of Scientific and Industrial Research for granting permission for publication. The authors are also indebted to Mr. F. M. Colebrook for the design and construction of the receiver used in the investigation, to Dr. R. L. Smith-Rose for advice on the method of presentation of the paper, and to Mr. B. J. Byrne for assistance in the experimental work.

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* Equation (2) was deduced on the assumption that the reflection coefficient of the ground was exactly − 1. A more complete expression for the field, taking into account the departure of the value of the reflection coefficient from unity, is

$$E = \frac{120\pi^2 AI}{\lambda^2 d}\sqrt{\left[\left(\frac{4\pi h_1 h_2}{\lambda d}\right)^2 + \frac{4(h_1 + h_2)^2}{(\kappa - 1) - 2j\frac{\sigma}{f}}\right]}$$

in which κ and σ are the dielectric constant and conductivity of the ground at the frequency f, the other symbols having the same meaning as in equation (2). This more exact expression for the field suggests that for the conditions in the present experiments the calibrations using equation (2) is approximately 20 % too small. The results of the experiment with varying height of transmitter at a fixed distance, however, suggest that the required correction was smaller than this. The exact amount, if any, of this correction was impossible to determine with the receiver used in the tests.

THE APPLICATION OF TRANSMISSION-LINE THEORY TO CLOSED AERIALS*

By F. M. COLEBROOK, B.Sc.

[From the National Physical Laboratory.]

(*Paper first received* 22*nd December*, 1937, *and in final form* 26*th February*, 1938.)

SUMMARY

The paper contains an analysis of the behaviour of frame aerials consisting of a single turn of conductor. It is assumed that the behaviour of such systems, including the mutual interactions of the various elements, can, to a useful degree of approximation, be represented by the differential equations of classical transmission-line theory. Formulae are obtained for the effective induced e.m.f. and the effective impedance (at the tuning point) of frame aerials the dimensions of which are not small compared with the wavelength, both for symmetrical and asymmetrical systems of tuning. It is found that in the case of symmetrically tuned systems the output voltages across the two equal halves of the tuning impedance will not in general be quite equal.

It is shown that the "resonance factor" of a frame aerial (i.e. the ratio of output voltage to induced e.m.f.) can be determined by the usual method of reactance-variation at a constant frequency, in spite of the non-uniformity of the current distribution along the length of the conductor, but that the same process carried out by variation of frequency will not, in general, be valid.

It is shown that for a given total length of conductor the optimum shape of a rectangular frame aerial, with respect to induced e.m.f., is square. In particular, a square frame with side equal to half a wavelength appears to have useful practical characteristics in respect of sensitivity and symmetry, both for field-strength measurement and direction-finding.

The method of applying the formulae to circular loops by a process of integration is given and illustrated by particular cases. It is found that in the case of small closed aerials the magnitude of the induced e.m.f. is not very sensitive to shape.

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(7) Physical Interpretation of Formulae.
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(1) OBJECT AND SCOPE

The most important uses of closed-coil or "frame" aerials are (*a*) directional reception and (*b*) field-strength measurement. The present paper has some bearing on the first of these, but is chiefly concerned with the second. In this application the e.m.f. induced in a closed aerial of total area turns A, located with its plane perpendicular to the wave-front of a linearly polarized electric wave of intensity e, is assumed to be given by

$$E = \frac{2\pi}{\lambda} Ae \qquad . \quad . \quad . \quad . \quad . \quad . \quad (1.1)$$

Further, if the closed aerial have total inductance L_a and total effective resistance R_a, the potential difference V produced across the tuning condenser is given by

$$V = \frac{\omega L_a}{R_a} E \qquad . \quad . \quad . \quad . \quad . \quad . \quad (1.2)$$

$$= \frac{\omega L_a}{R_a} \cdot \frac{2\pi}{\lambda} Ae \qquad . \quad . \quad . \quad . \quad . \quad (1.3)$$

Thus, to borrow a term which has come into use in direction-finding, the "pick-up factor" of the closed aerial is

$$\frac{V}{e} = \frac{\omega L_a}{R_a} \cdot \frac{2\pi A}{\lambda} \qquad . \quad . \quad . \quad . \quad . \quad (1.4)$$

consisting of two terms, a "resonance factor" and an effective e.m.f. factor.

The above simple formulations are based on the assumption that the total length of the conductor and all other dimensions of the closed aerial are so small compared with the wavelength that the current distribution is substantially uniform along the length of the conductor.

In many practical cases, particularly in the measurement of field strength in the metre band of wavelengths,

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* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 403.

the limitations of physical size involved in these assumptions are inconvenient. The principal object of the present paper is to arrive at formulae corresponding to those given for V and E above, which shall be valid for cases in which the current distribution cannot be assumed to be uniform, or the linear dimensions small compared with the wavelength. A secondary object was to examine the possibilities of certain special sizes and shapes of single-turn closed aerials, i.e. cases in which the principal dimensions are simply related to the wavelength of operation.

The present paper is purely analytical in character. Experimental work on the subject is in progress, and it is hoped to publish some account of this in the near future. Meanwhile, however, it has been thought desirable to publish the present theoretical discussion, since field-strength measurement, particularly at short wavelengths, is a subject of growing importance which is already being studied by a number of different groups of workers. The formulae developed in the analysis may be of some guidance in the interpretation of existing and accumulating experimental data. These data, on the other hand, may throw some light on the validity of the assumptions on which the analysis is based. Thus the theoretical conclusions are put forward in this tentative sense, both as furnishing suggestion for experimental work and as material for criticism in the light of practical experience.

(2) THE METHOD OF ANALYSIS AND THE ASSUMPTIONS INVOLVED

In a previous paper* it has been shown that the application of the classical transmission-line theory to open earthed aerials leads to theoretical conclusions which are, to a practically useful degree, in accordance with the actual behaviour of such aerials. Essentially, the physical assumption involved is that the linear conductor constituting the aerial can be regarded as having, at any given frequency, a uniformly distributed resistance, inductance, and capacitance, per unit length.

It might be thought that the existence of radiation resistance in aerials would invalidate such an assumption from the outset, but this is not necessarily the case. In fact, Brainerd has shown that† radiation resistance, or "radiactance" as he prefers to call it, can be represented as an addition, to the ohmic resistance per unit length, of a term consisting of a constant multiplied by the square of the frequency. The proposed formulation may therefore be valid at any given frequency, inclusive of radiation resistance.

Again, there will undoubtedly be some reaction of one part of a closed aerial on the remainder, but even this will not necessarily invalidate the proposed formulation, since such reaction may conceivably be represented for the most part by an effect on the magnitude of the assumed constants rather than a change of the form of the differential equations. Some such explanation must exist for the satisfactory agreement already mentioned between theory so based and experimental observation in the case of the open earthed aerial.

Thus there are good reasons for trying out the proposed formulation in the case of closed aerials, at least in the case of closed aerials consisting of a single turn, such as are generally used at short wavelengths. The formulae so obtained are likely to be somewhat more accurate than those which do not take into account the non-uniformity of the current distribution.

(3) THE BASIC SYSTEM CONSIDERED

The basic subject of analysis is illustrated in Fig. 1 (the closed loop of conductor is shown as circular in shape but is not necessarily assumed to be so in fact). The length of the conductor (l) is considered as divided into three parts l_1, l_2, and l_3, in the second of which a uniform e.m.f. e per unit length is induced by the field. All the practical cases involved can be derived as special cases or combinations of such systems.

As usual in single transmission-line theory, the potentials along the conductor are relative to earth as zero. For generality, two tuning impedances are shown, connected in series. They may in practice be equal, as in so-called balanced or symmetrical systems, or either may be zero. In general one point on these tuning impedances (e.g. the mid-point in a balanced system) will be connected either directly to earth or to the "earthy" point of a receiver or other measuring equipment. There may be, and in general will be, current in any such connection. Moreover, this point cannot in general be assumed to be at earth potential. It will, indeed, appear in the analysis that even if the frame is isolated at all points from earth the mid-point is not, in general, a node of potential, and some finite value must therefore be assumed for its potential. Again, if connected to earth, there may be, and generally will be, an e.m.f. induced in the earth lead by the incident field. The "earthed" point is thus assumed to be at a potential v_g above earth, and a current i_g is assumed to flow in the lead or connection to the "earthed" point.

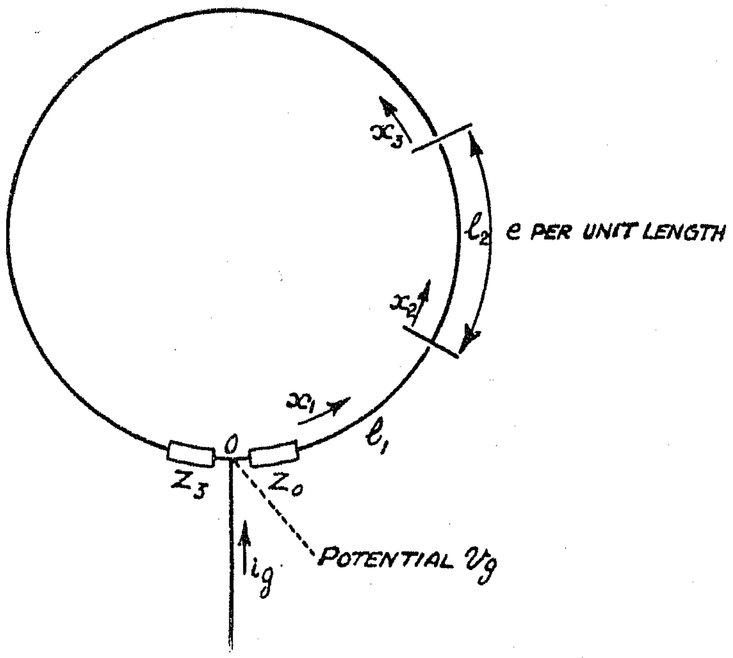

Fig. 1

As explained in Section (2), the conductor is assumed to have a uniform R, L, and C, per unit length. The current co-ordinates x, with suffixes 1, 2, 3, refer to distances along l_1, l_2, l_3, respectively. The instantaneous values of current i and potential v will be similarly distinguished. In addition, suffixes 0 and 1 will denote initial and final boundary values, i.e. i_{20} and i_{21} are values of i_2 at $x_2 = 0$ and $x_2 = l_2$.

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* See Reference (1).

† *Ibid.*, (2).

The symbols i and v are to be interpreted as rotating vectors, and e as a rotating vector $e\epsilon^{j\omega t}$. Any constants in the solution will thus have the character of vectors or of vector operators.

(4) ANALYSIS OF THE BASIC SYSTEM

For the parts l_1 and l_3 the appropriate differential equations are

$$(R + j\omega L)i = -\frac{\partial v}{\partial x} \quad . \quad . \quad . \quad (4.01)$$

$$j\omega Cv = -\frac{\partial i}{\partial x} \quad . \quad . \quad . \quad (4.02)$$

with suffixes 1 and 3. For the part l_2 it is necessary to include the e.m.f. e per unit length, giving

$$(R + j\omega L)i_2 = -\frac{\partial v_2}{\partial x_2} + e \quad . \quad . \quad (4.03)$$

and

$$j\omega Cv_2 = -\frac{\partial i_2}{\partial x_2} \quad . \quad . \quad . \quad (4.04)$$

The solutions of these are well known and can be written

$$i = A \sinh Px + B \cosh Px \quad . \quad . \quad . \quad (4.05)$$

$$v = -Z(B \sinh Px + A \cosh Px) \quad . \quad . \quad (4.06)$$

with suffixes 1 and 3 for l_1 and l_3, and

$$i_2 = A_2 \sinh Px_2 + B_2 \cosh Px_2 + \frac{e}{PZ} \quad . \quad (4.07)$$

$$v_2 = -Z(B_2 \sinh Px_2 + A_2 \cosh Px_2) \quad . \quad (4.08)$$

for l_2. In the above

$$P^2 = (R + j\omega L)j\omega C \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4.09)$$

$$Z^2 = \frac{R + j\omega L}{j\omega C} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4.10)$$

For the evaluation of the six unknown constant vectors A, B, there are the following boundary conditions:

$$\left.\begin{aligned} v_{10} &= -Z_0 i_{10} + v_g \\ v_{20} &= v_{11} \\ v_{30} &= v_{21} \\ v_{31} &= Z_3 i_{31} + v_g \\ i_{11} &= i_{20} \\ i_{21} &= i_{30} \end{aligned}\right\} \quad . \quad . \quad . \quad (4.11)$$

The detailed analysis, though lengthy, is quite straightforward and need not be given in full. The results obtained are:—

$$i_0 \frac{\sinh P(a_0 + a_3 + l)}{\sinh Pa_0} = \frac{2e}{PZ} \sinh \frac{Pl_2}{2} \sinh P\left(a_3 + l_3 + \frac{l_2}{2}\right)$$
$$+ \frac{2v_g}{Z} \sinh \frac{Pl}{2} \cosh P\left(a_3 + \frac{l}{2}\right) \quad . \quad (4.12)$$

and

$$i_3 \frac{\sinh P(a_0 + a_3 + l)}{\sinh Pa_3} = \frac{2e}{PZ} \sinh \frac{Pl_2}{2} \sinh P\left(a_0 + l_1 + \frac{l_2}{2}\right)$$
$$- \frac{2v_g}{Z} \sinh \frac{Pl}{2} \cosh P\left(a_0 + \frac{l}{2}\right) \quad . \quad (4.13)$$

In the above, l is the total length of the conductor, and i_0 is written for i_{10} and i_3 for i_{31}, that is, i_0 and i_3 are the currents through Z_0 and Z_3 respectively. Also the impedances Z_0 and Z_3 are expressed as equivalent real or complex lengths of line a_0 and a_3 by the transformations

$$Z_0 = Z \coth Pa_0$$

and

$$Z_3 = Z \coth Pa_3 \quad . \quad . \quad . \quad (4.14)$$

All the formulae appropriate to rectangular or circular loops, with balanced or unbalanced tuning, can be derived, by suitable specialization, from equations (4.12) and (4.13).

It will be demonstrated that the simplest and most practical short-wave closed-aerial system is that in which complete circuit symmetry is maintained, and Sections (5) to (13) are devoted to this system.

(5) SYMMETRICAL SYSTEMS

The symmetrical or balanced system is defined by

$$Z_3 = Z_0$$

i.e.

$$a_3 = a_0 \quad . \quad . \quad . \quad . \quad (5.1)$$

The basic equations become

$$i_0 \frac{\sinh P(2a_0 + l)}{\sinh Pa_0} = \frac{2e}{PZ} \sinh \frac{Pl_2}{2} \sinh P\left(a_0 + l_3 + \frac{l_2}{2}\right)$$
$$+ \frac{2}{Z}v_g \sinh \frac{Pl}{2} \cosh P\left(a_0 + \frac{l}{2}\right) \quad . \quad (5.2)$$

and

$$i_3 \frac{\sinh P(2a_0 + l)}{\sinh Pa_0} = \frac{2e}{PZ} \sinh \frac{Pl_2}{2} \sinh P\left(a_0 + l_1 + \frac{l_2}{2}\right)$$
$$- \frac{2}{Z}v_g \sinh \frac{Pl}{2} \cosh P\left(a_0 + \frac{l}{2}\right) \quad . \quad (5.3)$$

(6) RECTANGULAR FRAME: BALANCED TUNING

Consider the system illustrated in Fig. 2. Let i_0, i_3 and i'_0, i'_3, be the components of current due to the intensities $e\epsilon^{-j\theta}$ and $-e\epsilon^{j\theta}$ respectively, where $\theta = 2\pi b/\lambda$. These currents can be determined directly from (5.2) and (5.3) by appropriate substitutions for l_1, l_2, and l_3, in terms of the height h and width $2b$ of the frame, i.e.

$$l_1 = b,\ l_2 = h, \text{ and } l_3 = 3b + h.$$

Hence

$$l_3 + \frac{l_2}{2} = 3b + 3\frac{h}{2} = \frac{3l}{4}$$

and

$$l_1 + \frac{l_2}{2} = b + \frac{h}{2} = \frac{l}{4} \quad . \quad . \quad . \quad . \quad (6.01)$$

for the intensity $+e\epsilon^{-j\theta}$, and similarly for $-e\epsilon^{j\theta}$. Thus

$$i_0 = A\epsilon^{-j\theta} \sinh P\left(a_0 + \frac{3l}{4}\right) + Bv_g \quad . \quad . \quad (6.02)$$

$$i'_0 = -A\epsilon^{j\theta} \sinh P\left(a_0 + \frac{l}{4}\right) + Bv_g \quad . \quad . \quad (6.03)$$

$$i_3 = A\epsilon^{-j\theta} \sinh P\left(a_0 + \frac{l}{4}\right) - Bv_g \quad . \quad . \quad (6.04)$$

$$i'_3 = -A\epsilon^{j\theta} \sinh P\left(a_0 + \frac{3l}{4}\right) - Bv_g \quad . \quad . \quad (6.05)$$

where A and B are constant vector operators given by

$$A = \frac{2e}{PZ}\sinh\frac{Ph}{2}\Big/\left[\frac{\sinh P(2a_0+l)}{\sinh Pa_0}\right] \quad . \quad . \quad . \quad (6.06)$$

and

$$B = \frac{2}{Z}\sinh\frac{Pl}{2}\cosh P\left(a_0+\frac{l}{2}\right)\Big/\left[\frac{\sinh P(2a_0+l)}{\sinh Pa_0}\right] \quad (6.07)$$

When the intensities act simultaneously on the frame, the total currents will be given by the addition of the formulae in pairs. (It must be remembered, however, that v_g is a quantity, at present undetermined, dependent on i_0, i_0', i_3, and i_3', and in the addition of these formulae the terms in v_g are not added but merely replaced by a similar symbol for subsequent determination.)

The Asymmetry of the Balanced Frame Aerial

One immediate conclusion can be drawn from the general form of equations (6.02) to (6.05), namely that the balanced frame aerial is electrically unsymmetrical in general, in the sense that the potential differences across the two equal halves of the tuning impedance are not in general equal in magnitude and phase. The possibility of this was of course implied in the assumption of a current in the connection to the mid-point of the tuning impedance, but it remains to be shown that this current will in general exist.

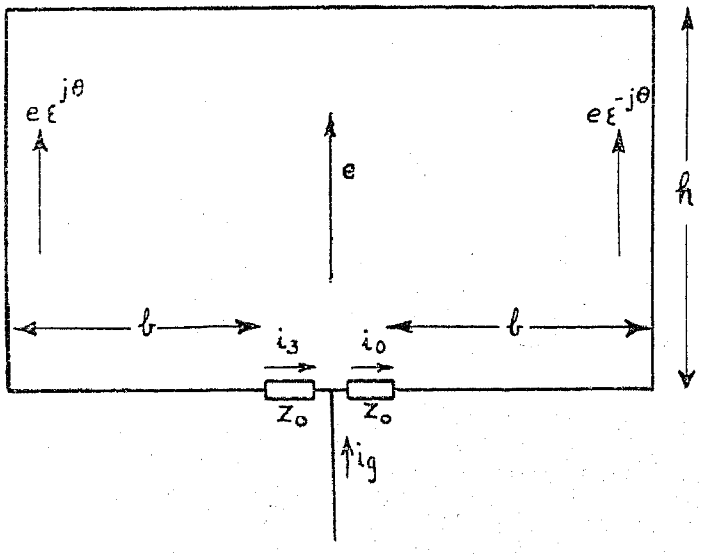

Fig. 2

Suppose, for example, that the mid-point is actually connected to earth by a lead so short that its impedance and any e.m.f. induced in it by the field can be neglected. Analytically, this is equivalent to putting $v_g = 0$ in equations (6.02) to (6.05). The circuit system is still completely balanced, but there will nevertheless be a current in this earth lead and a consequent small inequality of the potential differences across the two halves of the tuning impedance.

The general character of the vector system represented by i_0, i_0', i_3, and i_3', is illustrated in Fig. 3(*a*), which shows that the vector sums $(i_0 + i_0')$ and $(i_3 + i_3')$ are equal in magnitude but differ in phase. Thus in general the potential difference across the whole tuning impedance will not be twice the potential difference across either half of it. This asymmetry must obviously be a consequence of the fact that the complete system of circuit and field is not in fact symmetrical so far as a circulation round the loop is concerned. This asymmetry is able to manifest itself at the tuning impedances. In fact the different arguments of the hyperbolic functions in, for example, (6.02) and (6.03) correspond to the fact that the individual e.m.f.'s are not symmetrically disposed in relation to the tuning impedances.

It will be shown later that when the loop is tuned

$$a_0 = \tfrac{1}{4}n\lambda - \tfrac{1}{2}l; \quad n = 1, 3, 5, \text{etc.} \quad . \quad (6.08)$$

Under these conditions

$$a_0 + \tfrac{1}{4}l = \tfrac{1}{4}n\lambda - \tfrac{1}{4}l$$

and

$$a_0 + \tfrac{3}{4}l = \tfrac{1}{4}n\lambda + \tfrac{1}{4}l \quad . \quad . \quad . \quad (6.09)$$

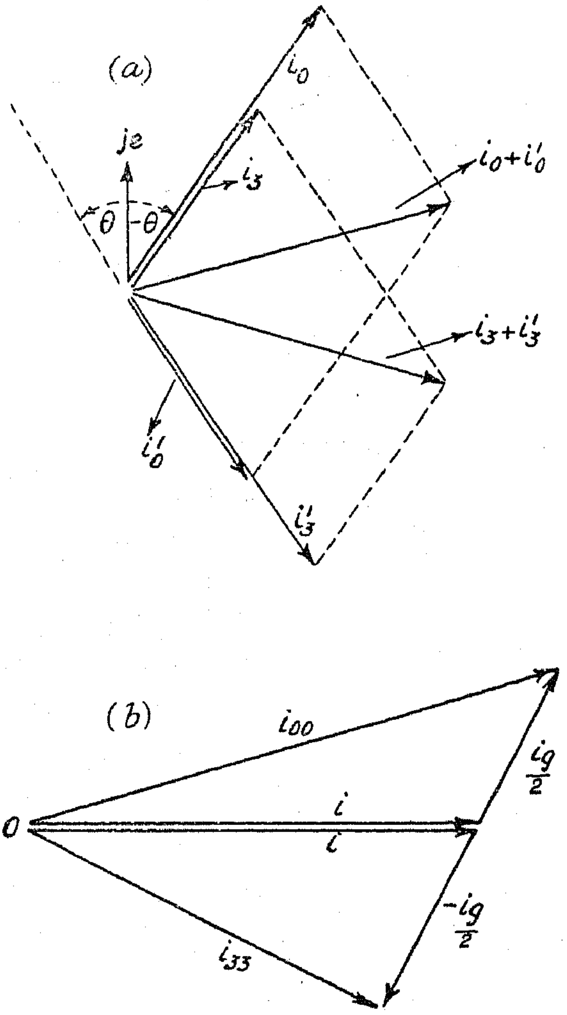

Fig. 3

and the sinh terms become equal in phase and magnitude except for small terms due to attenuation. Thus when the loop is tuned the currents $(i_0 + i_0')$ and $(i_3 + i_3')$ through the two tuning impedances become practically equal in phase and magnitude, except for a very small and probably negligible effect due to differences of attenuation in the lengths $\frac{1}{4}l$ and $\frac{3}{4}l$ of the conductor.

It is important to remember, however, that this is true only on the assumption that the only e.m.f.'s acting on the system are those induced in the loop by the incident field. It is true, that is to say, for a very short lead to earth or to the "earthy" point of a receiver. Suppose, however, there is a relatively long lead from the mid-point. Then there may be induced in this lead an e.m.f. which may be of the same order as, or even greater than, the e.m.f.'s induced in the sides of the loop, and there may be in consequence an appreciable current in the mid-point connection due to this e.m.f. If the loop is truly balanced this current will divide equally at the mid-point, adding to the current through one half and subtracting from the current through the other half. Thus we may have a state of affairs somewhat as shown in Fig. 3(*b*), and again the potentials across each half of the tuning impedance will be unequal, possibly both in magnitude and phase.

The effect of the mid-point current is of course eliminated if the output voltage be measured as the vector sum of the voltages across each half of the tuning impedance. This point is shown in analytical detail below, and the practical aspect of the matter is referred to again in Section (16). It may be pointed out, however, that these considerations do not apply exclusively to the balanced loop as a means of field-strength measurement. A balanced dipole system may be subject to the same considerations in respect of e.m.f.'s, if any, induced in a conductor connected to the mid-point of any tuning impedance associated with it.

The addition of (6.02), (6.03) and (6.04), (6.05) gives for the total currents $(i_0 + i_0')$ and $(i_3 + i_3')$ due to the simultaneous action of the e.m.f.'s induced in the two sides of the frame:—

$$i_0 + i_0' = A\left\{2\cosh P\left(a_0 + \frac{l}{2}\right)\sinh\frac{Pl}{4}\cos\theta - 2j\sinh P\left(a_0 + \frac{l}{2}\right)\cosh\frac{Pl}{4}\sin\theta\right\} + Bv_g \quad . \quad (6.10)$$

and

$$i_3 + i_3' = A\left\{-2\cosh P\left(a_0 + \frac{l}{2}\right)\sinh\frac{Pl}{4}\cos\theta - 2j\sinh P\left(a_0 + \frac{l}{2}\right)\cosh\frac{Pl}{4}\sin\theta\right\} - Bv_g \quad . \quad (6.11)$$

[In these equations, as already pointed out, v_g is as yet undetermined, and will not have the same value as in equations (6.02) to (6.05).]

The first point to note is that the sum of the combined currents (and therefore the total potential difference across the tuning impedance) is independent of v_g. This is otherwise obvious. It is not immediately obvious that it is also independent of i_g, the current in the "earth" connection, but this is clear if we assume, in accordance with the circuit symmetry, that the "earth" current divides equally at the mid-point. Writing i_{00} and i_{33} for the combined currents, then, as illustrated in Fig. 4,

$$i = \tfrac{1}{2}(i_{00} + i_{33}) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (6.12)$$

and

$$i_g = i_{00} - i_{33} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (6.13)$$

$$i_{00} = \tfrac{1}{2}(i_{00} + i_{33}) + \tfrac{1}{2}(i_{00} - i_{33}) = i + \tfrac{1}{2}i_g \quad (6.14)$$

and

$$i_{33} = \tfrac{1}{2}(i_{00} + i_{33}) - \tfrac{1}{2}(i_{00} - i_{33}) = i - \tfrac{1}{2}i_g \quad (6.15)$$

Thus i can be regarded as a circulating current flowing through the two halves of the impedance in series, while an "earth" current i_g flows through the two halves in parallel, $\frac{1}{2}i_g$ flowing through each half. From (6.10) and (6.11) we have

$$i = -2jA\sinh P\left(a_0 + \frac{l}{2}\right)\cosh\frac{Pl}{4}\sin\theta \quad . \quad (6.16)$$

and

$$i_g = 4A\cosh P\left(a_0 + \frac{l}{2}\right)\sinh\frac{Pl}{4}\cos\theta + 2Bv_g \quad (6.17)$$

From (6.17) it can be seen, as already stated, that i_g is not zero in general even when v_g is taken to be zero, and further that, apart from any variation of v_g, a value of θ which makes i a maximum will make i_g a minimum.

Inserting in (6.16) and (6.17) the values of A and B from (6.06) and (6.07) gives

$$i\,\frac{\cosh P(a_0 + \frac{1}{2}l)}{\sinh Pa_0} = \frac{-2je}{PZ}\sinh\frac{Ph}{2}\cosh\frac{Pl}{4}\sin\theta \quad (6.18)$$

and

$$i_g\,\frac{\sinh P(a_0 + \frac{1}{2}l)}{\sinh P\,a_0} = \frac{4e}{PZ}\sinh\frac{Ph}{2}\sinh\frac{Pl}{4}\cos\theta + \frac{2v_g}{Z}\sinh\frac{Pl}{2} \quad . \quad . \quad . \quad (6.19)$$

It is shown in the next Section that these formulae admit of a simple physical interpretation.

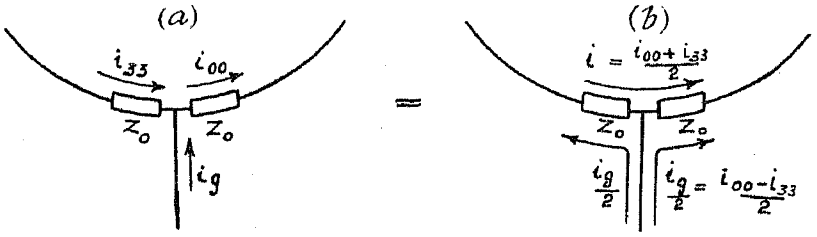
(a)
i_{33}
i_{00}
Z_0
Z_0
i_g
=
(b)
$i = \frac{i_{00}+i_{33}}{2}$
Z_0
Z_0
$\frac{i_g}{2}$
$\frac{i_g}{2} = \frac{i_{00}-i_{33}}{2}$

Fig. 4

(7) PHYSICAL INTERPRETATION OF FORMULAE

Formula (6.18) admits of a very simple physical interpretation. Replacing coth Pa_0 by Z_0/Z and θ by $2\pi b/\lambda$ gives

$$i\left\{2Z_0 + 2Z\tanh\frac{Pl}{2}\right\} = \frac{-4je}{P}\,\frac{\cosh(\frac{1}{4}Pl)\sinh(\frac{1}{2}Ph)\sin(2\pi b/\lambda)}{\cosh(\frac{1}{2}Pl)} \quad (7.01)$$

or

$$i = \frac{e_e}{2Z_0 + Z_e} \quad . \quad . \quad . \quad . \quad (7.02)$$

where

$$e_e = \frac{-4je}{P}\,\frac{\cosh(\frac{1}{4}Pl)\sinh(\frac{1}{2}Ph)\sin(2\pi b/\lambda)}{\cosh(\frac{1}{2}Pl)} \quad (7.03)$$

and

$$Z_e = 2Z\tanh(\tfrac{1}{2}Pl) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7.04)$$

Thus the circulating current i can be regarded as due to an effective e.m.f. e_e acting in a circuit consisting of the tuning impedance $2Z_0$ in series with an effective aerial impedance Z_e, where both e_e and Z_e are quite independent of the tuning impedance.

It will be shown, moreover, that if the loop is considered to be small in dimensions compared with λ, these formulae reduce to those given in the first section for small aerials. Thus, when l and h are small compared with λ,

$$\cosh(\tfrac{1}{2}Pl) \rightarrow \cosh(\tfrac{1}{4}Pl) \rightarrow 1$$

$$\sinh(\tfrac{1}{2}Ph) \rightarrow \tfrac{1}{2}Ph$$

and

$$\sin\frac{2\pi b}{\lambda} \rightarrow \frac{2\pi b}{\lambda} \quad . \quad . \quad . \quad . \quad (7.05)$$

and
$$e_e \to -\frac{4je}{P}\cdot\frac{Ph}{2}\cdot\frac{2\pi b}{\lambda}$$

$$= -j\frac{2\pi}{\lambda}Ae \quad . \quad . \quad . \quad . \quad . \quad . \quad (7.06)$$

i.e.
$$|e_e| = \frac{2\pi}{\lambda}A|e| \quad . \quad . \quad . \quad . \quad . \quad . \quad (7.07)$$

where $A = 2bh =$ area of loop.

Further,
$$2Z\tanh(\tfrac{1}{2}Pl) \to PZl$$
$$= (R + j\omega L)l$$
$$= R_a + j\omega L_a \quad . \quad . \quad . \quad . \quad (7.08)$$

where R_a and L_a are the total resistance and inductance of the loop. When the aerial is tuned

$$2Z_0 + j\omega L_a = 0 \quad . \quad . \quad . \quad (7.09)$$

i.e.
$$i = \frac{-j\frac{2\pi}{\lambda}Ae}{R_a} \quad . \quad . \quad . \quad . \quad . \quad . \quad (7.10)$$

and
$$v = 2Z_0 i = -j\omega L_a i$$
$$= -\frac{\omega L_a}{R_a}\cdot\frac{2\pi}{\lambda}\cdot A.e$$

or
$$|v| = \frac{\omega L_a}{R_a}|e_e| \quad . \quad . \quad . \quad . \quad . \quad . \quad (7.11)$$

The formula for the earth current i_g admits of a similar interpretation. The same process gives

$$i_g\left\{\frac{Z}{2}\coth\frac{Pl}{2} + \frac{Z_0}{2}\right\} = \frac{e}{P}\cdot\frac{\sinh(Ph/2)}{\cosh(Pl/4)}\cos\frac{2\pi b}{\lambda} + v_g \quad (7.12)$$

In this case the effective aerial impedance is $\frac{1}{2}(Z\coth\frac{1}{2}Pl)$ (i.e. two paths in parallel) and the added impedance is that of two impedances Z_0 in parallel, while the effective e.m.f. is a term due to e, plus the assumed potential of the mid-point, i.e. v_g.

(8) DETERMINATION OF THE MID-POINT POTENTIAL AND EARTH-LEAD CURRENT

If, in the case illustrated in Fig. 2, the earth lead, of length g, is assumed to have the same value of P as the conductor forming the frame aerial, is earthed at its lowest point, and is vertical and subject to a field intensity e per unit length, it is easily shown, by the methods used in Section (4), that

$$v_g = -Z\tanh Pg\left(i_g - \frac{e}{PZ}\right) \quad . \quad . \quad (8.1)$$

In combination with (7.12) this gives, for i_g,

$$i_g\left(\frac{Z_0}{2} + \frac{Z}{2}\coth\frac{Pl}{2} + Z\tanh Pg\right)$$
$$= \frac{e}{P}\left(\frac{\sinh(Ph/2)}{\cosh(Pl/4)}\right)\cos\frac{2\pi b}{\lambda} + \tanh Pg \quad (8.2)$$

Alternatively, the mid-point may not be connected to earth directly. It might, for example, be connected by a short lead to the "earthy" point of a receiver, of which the capacitance to earth can be represented as an impedance Z_g.

In this case

$$v_g = -Z_g i_g \quad . \quad . \quad . \quad . \quad (8.3)$$

and

$$i_g\left(\frac{Z_0}{2} + \frac{Z}{2}\coth\frac{Pl}{2} + Z_g\right) = \frac{e}{P}\cdot\frac{\sinh(Ph/2)}{\cosh(Pl/4)}\cos\frac{2\pi b}{\lambda} \quad (8.4)$$

Even in this case, however, i_g will not necessarily be negligible, since the combined impedance term on the left of (8.4) is clearly capable of resonance conditions, as also is the e.m.f. on the right of (8.4).

Finally, as a point of academic interest, if it be assumed that the frame can be so disposed that there is no current from the mid-point to earth, i.e. if there is no connection from the mid-point to earth, and any apparatus connected to the mid-point has negligible capacitance to earth, then, from (7.12),

$$v_g = -\frac{e}{P}\frac{\sinh(Ph/2)}{\cosh(Pl/4)}\cos\frac{2\pi b}{\lambda} \quad . \quad . \quad (8.5)$$

The formula for i is not affected, but it will be found that the substitution of this value for v_g in the formulae for i_{00} and i_{33} makes these currents equal in phase and magnitude, which is otherwise obvious.

The physical significance of the finite value of v_g given by (8.5) is that the mid-point of a balanced frame aerial completely isolated from earth is not in general a node of potential, though, from (8.5), it becomes a node of potential if the width $2b$ is an odd number of half wavelengths. This is in agreement with the conclusions of Palmer, Taylor, and Witty, based on considerations of current distribution.*

(9) THE MEASUREMENT OF THE RESONANCE FACTOR

The term "resonance factor" is here used, for want of any existing and generally recognized term, to denote the ratio between v, the potential difference across the tuning circuit, and e_e, the effective e.m.f. induced in the loop. It is thus a generalization of the term Q or "magnification factor" as applied to uniform-current circuits.

In the case of closed-coil aerials small compared with the wavelength, this factor is usually determined by one of two methods, (*a*) reactance variation, (*b*) insertion of a small known e.m.f. and measurement of the corresponding terminal potential difference.

It will be desirable to determine whether these or similar methods are valid for cases in which the coil aerial is not small compared with the wavelength, and in which therefore the current distribution is not uniform.

(a) Reactance Variation

It has been shown in Section (7) that for a rectangular frame aerial the circulating current i can be regarded as due to an effective e.m.f. e_e acting in series with a variable tuning impedance $2Z_0$ and an aerial impedance Z_e, i.e.

$$i = \frac{e_e}{2Z_0 + Z_e} \quad . \quad . \quad . \quad . \quad (9.01)$$

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* *Proceedings of the Physical Society*, 1934, vol. 46, p. 76.

In this expression e_e and Z_e are independent of Z_0, but both depend on frequency. At any given frequency, however, Z_e can be regarded as consisting of an effective resistance R_e in series with an effective reactance X_e. The tuning impedance may take the form of a symmetrical condenser, and if C_0 be the magnitude of the two halves in series

$$2Z_0 = \frac{1}{j\omega C_0} \quad . \quad . \quad . \quad . \quad . \quad (9.02)$$

The exact tuning condition for resonance of the potential difference v across C_0 is given by

$$\omega C_0 = \frac{X_e}{Z_e^2} \quad . \quad . \quad . \quad . \quad . \quad (9.03)$$

where $$Z_e^2 = R_e^2 + X_e^2 \quad . \quad . \quad . \quad (9.04)$$

and at resonance $$\left|\frac{v}{e_e}\right|^2 = \frac{Z_e^2}{R_e^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (9.05)$$

If C_0 be adjusted to a new value C_0' for which the condenser potential difference is reduced to $1/\sqrt{2}$ of its resonance-value, then

$$\frac{1}{Z_e^2}\,\frac{1}{\left(\frac{R_e}{Z_e^2}\right)^2 + \left(\omega C_0' - \frac{X_e}{Z_e^2}\right)^2} = \frac{1}{2}\cdot\frac{Z_e^2}{R_e^2} \quad . \quad (9.06)$$

whence $$\left(\omega C_0' - \frac{X_e}{Z_e^2}\right)^2 = \frac{R_e^2}{Z_e^4} \quad . \quad . \quad . \quad (9.07)$$

or $$\omega C_0' - \frac{X_e}{Z_e^2} = \pm\frac{R_e}{Z_e^2} \quad . \quad . \quad (9.08)$$

If C_0' and C_0'' be the two values of C_0 which satisfy this condition

$$\omega C_0' - \frac{X_e}{Z_e^2} = \frac{R_e}{Z_e^2} \quad . \quad . \quad . \quad . \quad . \quad (9.09)$$

and $$\omega C_0'' - \frac{X_e}{Z_e^2} = -\frac{R_e}{Z_e^2} \quad . \quad . \quad . \quad . \quad (9.10)$$

or, putting

$$C_0' - C_0'' = 2\delta C_0 \quad . \quad . \quad . \quad (9.11)$$

$$\omega\delta C_0 = \frac{R_e}{Z_e^2} \quad . \quad . \quad . \quad . \quad . \quad (9.12)$$

Also $$\omega C_0 = \frac{X_e}{Z_e^2} \quad . \quad . \quad . \quad . \quad . \quad (9.13)$$

The combination of (9.05), (9.12), and (9.13), gives

$$\left|\frac{v}{e_e}\right| = \frac{C_0}{\delta C_0}\left(1 + \frac{\delta C_0^2}{C_0^2}\right)^{\frac{1}{2}} \quad . \quad . \quad . \quad (9.14)$$

In general δC_0^2 will be quite negligible compared with C_0^2, and

$$\left|\frac{v}{e_e}\right| = \frac{C_0}{\delta C_0} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (9.15)$$

Thus the resonance variation method is valid in spite of the non-uniform current distribution, provided it be carried out by variation of the tuning capacitance *and not by variation of frequency.* A formula appropriate to variation of frequency could probably be determined, but it would be very complicated and probably less accurate.

(b) Determination by Inserted E.M.F.

The validity of this method can be examined by reference to (5.12), i.e.

$$i\frac{\cosh P(a_0 + \frac{1}{2}l)}{\sinh Pa_0} = \frac{e}{PZ}\cosh\frac{P(l_3 - l_1)}{2}\sinh\frac{Pl_2}{2}$$

which can be put in the form

$$i = \frac{e_e}{Z_0 + Z_e}$$

where $$e_e = \frac{\frac{2e}{P}\sinh\left(\frac{1}{2}Pl_2\right)\cosh\left[\frac{1}{2}P(l_3 - l_1)\right]}{\left(\cosh\frac{1}{2}Pl\right)} \quad . \quad (9.16)$$

Suppose now that the inserted e.m.f. e_0 be located at the centre of the loop remote from the tuning circuit. This condition is represented by reducing the length l_2 to zero with the condition

$$\underset{l_2 \to 0}{lt.}\ \frac{2e}{P}\sinh\frac{Pl_2}{2} = \underset{l_2 \to 0}{lt.}\ el_2 = e_0 \quad . \quad (9.17)$$

Then, since for this condition $l_1 = l_3$,

$$e_e = \frac{e_0}{\cosh\left(\frac{1}{2}Pl\right)} \quad . \quad . \quad . \quad (9.18)$$

Thus the effective series e.m.f. is greater than e_0. For example, if $l = \frac{1}{5}\lambda$, $\cosh\left(\frac{1}{2}Pl\right) \simeq 0{\cdot}80$ and the resonance factor would be over-estimated by some 20 % if this method were used, and for $l = \frac{1}{4}\lambda$ the error would be 30%.

However, by applying the same process it is easy to show that if the e.m.f. be inserted adjacent to the tuning reactance, i.e. $l_1 = 0$, $l_3 = l$, then the effective series e.m.f. is equal to e_0. *Thus the inserted-e.m.f. method can be used, regardless of the relation of l to λ, provided the e.m.f. be inserted adjacent to the tuning circuit and not at the centre of the loop remote from the tuning circuit.* (This assumes that the output voltage is measured across the tuning circuit. The case, which sometimes occurs in practice, when the loop is tuned by a condenser in the top centre of the loop, and the output is measured across a fixed impedance at the bottom centre, would need separate consideration on the lines indicated.)

(10) DETAILED ANALYSIS OF RECTANGULAR FRAME WITH BALANCED TUNING

The following relationships are listed for reference.

$$P = \sqrt{\left[(R + j\omega L)j\omega C\right]}$$

$$= \alpha + j\beta$$

where, assuming R^4 is small compared with $\omega^4 L^4$,

$$\alpha = \frac{R}{2}\sqrt{\frac{C}{L}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10.01)$$

$$\beta = \omega\sqrt{(LC)}\left(1 + \frac{R^2}{8\omega^2 L^2}\right) \quad . \quad . \quad . \quad (10.02)$$

Therefore, neglecting $R^2/(8\omega^2L^2)$ compared with 1,

$$\beta = \omega\sqrt{(LC)} = \frac{2\pi}{\lambda} \quad . \quad . \quad . \quad . \quad . \quad (10.03)$$

Also

$$Z = \sqrt{\frac{L}{C}}\left\{\left(1 + \frac{R^2}{8\omega^2L^2}\right) - j\frac{R}{2\omega L}\right\} \quad . \quad (10.04)$$

$$\simeq \sqrt{\frac{L}{C}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10.05)$$

For the type of conductor contemplated in this analysis, the "characteristic impedance" Z will usually be about 600 ohms.

In the following analysis it is assumed that $\beta = 2\pi/\lambda$ and that $2\alpha\lambda$ is so small that $\cosh 2\alpha\lambda \simeq 1$ and $\sinh 2\alpha\lambda \simeq 2\alpha\lambda$.

The detailed examination of the rectangular frame is most conveniently related to the formulae

$$e_e = \frac{-4je}{P}\frac{\cosh(\frac{1}{4}Pl)\sinh(\frac{1}{2}Ph)\sin 2\pi b/\lambda}{\cosh(\frac{1}{2}Pl)} \quad (10.06)$$

$$Z_e = 2Z\tanh(\tfrac{1}{2}Pl) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10.07)$$

and

$$i_g\left(\frac{Z_0}{2} + \frac{Z}{2}\coth\frac{Pl}{2} + Z\tanh P_g\right)$$

$$= \frac{e}{P}\left[\frac{\sinh(\frac{1}{2}Ph)}{\cosh(\frac{1}{4}Pl)}\cos 2\pi b/\lambda + \tanh P_g\right] \quad (10.08)$$

The formula for i_g will make it possible to estimate the difference between the potential differences across the two halves of the tuning impedance, which can be expressed fractionally in the form

$$\frac{v_0 - v_3}{v} = \frac{i_0 - i_3}{2i} = \frac{1}{2}\cdot\frac{i_g}{i} \quad . \quad . \quad (10.09)$$

(a) Small Aerial

For the common case in which l is small compared with λ, but not negligibly small, the ratio of the effective e.m.f. to that calculated by the simple formula given in Section (1) is very approximately

$$\left|\frac{e_e}{e_e'}\right| = \frac{\sin\pi h/\lambda}{\pi h/\lambda}\,\frac{\sin 2\pi b/\lambda}{2\pi b/\lambda}\,\frac{\cos\pi l/2\lambda}{\cos\pi l/\lambda} \quad . \quad (10.10)$$

Taking, for example, a square loop with

$$2b = h = \lambda/16;\ \text{i.e. } l = \lambda/4$$

$$\left|\frac{e_e}{e_e'}\right| = \left(\frac{\sin\pi/16}{\pi/16}\right)^2\frac{\cos\pi/8}{\cos\pi/4} = 1\cdot23 \quad . \quad (10.11)$$

In this case, therefore, the use of the simpler formula would lead to an over-estimation of field strength of about 23 %.

It is difficult to estimate the significance of i_g in such a case, but if it be assumed in the first instance that the frame is earthed at its mid-point,

$$\frac{i_g}{2i} = \frac{j}{2}\cdot\frac{\cosh(\frac{1}{2}Pl)}{\cosh^2(\frac{1}{4}Pl)}\cot\frac{2\pi b}{\lambda}\cdot\frac{Z_0 + Z\tanh(\frac{1}{2}Pl)}{Z_0 + Z\coth(\frac{1}{2}Pl)} \quad (10.12)$$

Now

$$Z_0 + Z\coth\frac{Pl}{2} \simeq \frac{1}{j\omega C_0} + \frac{2Z}{Pl} = \frac{1}{j\omega C_0} + \frac{2}{j\omega C_a} \quad (10.13)$$

where C_a is written for lC, i.e. it is the total aerial capacitance. Also

$$Z_0 + Z\tanh\frac{Pl}{2} \simeq \frac{1}{j\omega C_0} + \frac{R_a}{2}\frac{j\omega L_a}{2} \quad . \quad (10.14)$$

and, if the aerial is tuned,

$$\frac{1}{j\omega C_0} = \frac{1}{2}j\omega L_a \quad . \quad . \quad . \quad . \quad (10.15)$$

In general $2/(j\omega C_a)$ will be large compared with $1/(j\omega C_0)$.

Therefore

$$\frac{Z_0 + Z\tanh(\frac{1}{2}Pl)}{Z_0 + Z\coth(\frac{1}{2}Pl)} \simeq \frac{1}{4}j\omega C_a R_a \quad . \quad . \quad (10.16)$$

Thus for the numerical case considered

$$\frac{i_g}{2i} = -\frac{1}{8}\left(\frac{\cos\pi/4}{\cos^2\pi/8}\right)\cot\frac{\pi}{16}\omega C_a R_a$$

$$= -0\cdot52\,\omega C_a R_a \quad . \quad . \quad . \quad . \quad . \quad . \quad (10.17)$$

But

$$\omega^2 LC = \frac{4\pi^2}{\lambda^2}$$

Therefore for

$$l = \lambda/4, \quad \omega C_a = \frac{\pi^2}{4\omega L_a} \simeq \frac{10}{4}\cdot\frac{1}{\omega L_a}$$

Therefore

$$\frac{i_g}{2i} = -0\cdot52 \times \frac{10}{4}\left(\frac{R_a}{\omega L_a}\right) = -1\cdot3\frac{R_a}{\omega L_a} \quad (10.18)$$

Thus the unsymmetrical current is very small compared with the circulating current, and may be 1 or 2 per cent of it in magnitude. If the frame is raised above the surface of the earth, however, and the centre point is connected to earth by a lead, the effective e.m.f. producing the unsymmetrical current will be increased—roughly in the proportion of g to $\frac{1}{2}h$, without a proportionate increase of impedance, and it therefore appears that even with comparatively small earthed frame aerials it cannot safely be assumed that the total tuning-circuit potential difference can be determined by measurement across one side only of the symmetrical tuning system.

It would, however, appear to be permissible to use such a frame without direct connection to earth, and the asymmetry may in practice be reduced in this way, though at the very high frequencies assumed in this analysis the capacitance to earth of the measuring system will, in general, make it difficult to maintain a very high impedance to earth. In practice, if the receiver potential difference is determined by measurement across one half of the tuning impedance it will in general be advisable to compare the values corresponding to the two maxima (by rotation of the frame). Any substantial inequality observed may be the result of the asymmetry discussed above.

(b) Large Frame Aerials

Provided that the theory of the large frame aerial be known there would seem to be no need to restrict the frame dimensions as is the present practice in field-strength measurement. In fact, the use of a large frame may offer some advantages. It will therefore be of practical interest to consider cases where l may be comparable with or greater than the wavelength of operation.

(1) $l = 0$ to $l = \frac{1}{2}\lambda$.

Up to, but not including, $l = \frac{1}{2}\lambda$ there are no resonances of impedance or e.m.f. The effective impedance remains positive (inductive) in character and the system is tunable by a capacitance. The e.m.f. can be calculated by the formulae given.

(2) $l = \frac{1}{2}\lambda$.

$$Z_e \simeq \frac{8Z}{\alpha\lambda} \simeq \frac{8L_a}{C_aR_a} \quad . \quad . \quad . \quad . \quad (10.19)$$

Thus Z_e is very large and resistive. The appropriate tuning system would be a parallel-tuned circuit. The e.m.f. is approximately

$$e_e = \frac{-2\sqrt{2}}{\alpha\lambda}\left(\frac{2\pi}{\lambda}Ae\right)$$

$$= -2\sqrt{(2)}\frac{Z}{R_a}\left(\frac{2\pi}{\lambda}Ae\right) \quad . \quad . \quad (10.20)$$

and may thus be large compared with the e.m.f. calculated by the uniform current formula. Thus the half-wavelength loop may be considered for field strength measurement, though it may be difficult to realize a tuning impedance of sufficient magnitude to make the output or received potential difference a large fraction of the induced e.m.f.

(3) $l = \frac{3}{4}\lambda$.

$$Z_e \simeq -jZ \quad . \quad . \quad . \quad . \quad . \quad (10.21)$$

i.e. the impedance is fairly large and negative. The e.m.f. will depend on the ratio of $2b$ to h, but will not show any resonant conditions.

(4) $l = \lambda$.

$$Z_e \simeq Z\alpha\lambda \simeq \frac{R_a}{2} \quad . \quad . \quad . \quad . \quad (10.22)$$

Thus the impedance is small and resistive. However, the e.m.f. contains the term $\cosh \frac{1}{4}Pl$, and this reduces to the very small quantity $\frac{1}{4}\alpha\lambda$. Thus the induced e.m.f. will be very small and the configuration appears to have no practical value.

(5) $l = \frac{3}{2}\lambda$.

$$Z_e \simeq \frac{4}{3\alpha\lambda}2Z \simeq \frac{8}{3}\cdot\frac{L_a}{C_aR_a} \quad . \quad . \quad . \quad (10.23)$$

Thus the impedance is similar to that in the $\frac{1}{2}\lambda$ case.

Also $$e_e = j\frac{2\sqrt{2}}{3}\cdot\frac{Z}{R_a}\left(\frac{2e\lambda}{\pi}\sinh\frac{Ph}{2}\sin\frac{2\pi b}{\lambda}\right) \quad . \quad (10.24)$$

and is therefore large in general and depends on the values of b and h. Thus if $2b = \frac{1}{2}\lambda$ and $h = \frac{1}{4}\lambda$

$$e_e = -\frac{2}{3}\cdot\frac{Z}{R_a}\cdot\frac{2e\lambda}{\pi} \quad . \quad . \quad . \quad (10.25)$$

and is therefore large.

(6) $l = 2\lambda$.

$$Z_e \simeq 2\,Z\alpha\lambda \simeq R_a \quad . \quad . \quad . \quad . \quad (10.26)$$

The impedance is therefore small and resistive. The appropriate tuning system is a series-tuned circuit.

$$e_e \simeq \frac{2\lambda e}{\pi}\left(\sinh\frac{Ph}{2}\sin\frac{2\pi b}{\lambda}\right) \quad . \quad . \quad (10.27)$$

and therefore depends on the shape. If the coil is square, i.e. $2b = h = \frac{1}{2}\lambda$

$$e_e \simeq j\frac{2\lambda e}{\pi} = j\frac{le}{\pi} \quad . \quad . \quad . \quad . \quad (10.28)$$

This appears to be a possible configuration for field-strength measurement. For example, the effective e.m.f. would be about 25 times as large as that given by a small coil having a periphery of $\frac{1}{4}\lambda$, and 40 times as large as that given by a coil with a periphery of $\frac{1}{5}\lambda$.

(11) THE LARGE RECTANGULAR FRAME IN RELATION TO DIRECTION-FINDING

It has already been shown that the large frame may be expected to have the advantage of a relatively large "pick-up factor." In relation to direction-finding, however, the important quantity is the sharpness of the minimum. For a vertical wave incident at an angle ϕ to the normal to the plane of the frame

$$\left|e_e\right| = \left|\frac{4_e}{P}\cdot\frac{\cosh(\frac{1}{4}Pl)\sinh(\frac{1}{2}Ph)}{\cosh(\frac{1}{2}Pl)}\cdot\frac{\sin 2\pi b\sin\phi}{\lambda}\right| \quad . \quad (11.1)$$

The sharpness of the minima can be measured by $d|e_e|/d\phi$, i.e.

$$S = \frac{d|e_e|}{d\phi} = \frac{2\lambda e}{\pi}\cdot\frac{\cos 2a\pi\sin a\pi}{\cos 4a\pi}\cos(a\pi\sin\phi)a\pi\cos\phi \quad . \quad . \quad (11.2)$$

By a detailed analysis of this formula it can be shown that a large frame, in particular the square frame with half-wavelength sides, may be expected to give a much greater sharpness of minimum than a small frame. It is realized, however, that sharpness of minimum, though important, is not the only or even the most important feature of a direction-finding system. Nevertheless, the large frame would seem to be worth considering in relation to this application.

(12) OPTIMUM SHAPE OF RECTANGULAR FRAME WITH BALANCED TUNING

For a given total length, the effective e.m.f. depends on the shape of the frame, in virtue of the terms

$$\sinh\frac{Ph}{2}\sin\frac{2\pi}{\lambda}b \quad . \quad . \quad . \quad . \quad (12.1)$$

Putting $$b = \frac{1}{4}l - \frac{1}{2}h \quad . \quad . \quad . \quad . \quad (12.2)$$

and neglecting the attenuation component of P, these terms become

$$\sin\frac{2\pi}{\lambda}\cdot\frac{h}{2}\sin\frac{2\pi}{\lambda}(\tfrac{1}{4}l - \tfrac{1}{2}h) \quad . \quad . \quad (12.3)$$

i.e. $$\cos\frac{2\pi}{\lambda}(h - \tfrac{1}{4}l) - \cos\frac{2\pi}{\lambda}\cdot\frac{l}{4} \quad . \quad . \quad (12.4)$$

The critical values of h are therefore those for which

$$\sin\frac{2\pi}{\lambda}(h - \tfrac{1}{4}l) = 0 \quad . \quad . \quad . \quad (12.5)$$

i.e. $$\frac{2\pi}{\lambda}(h - \tfrac{1}{4}l) = n\pi; \quad n = 0, 1, 2, 3, \text{ etc.} \quad . \quad (12.6)$$

or $$h = \tfrac{1}{4}l + \tfrac{1}{2}n\lambda \quad . \quad . \quad . \quad . \quad . \quad (12.7)$$

and $$2b = \tfrac{1}{4}l - \tfrac{1}{2}n\lambda \quad . \quad . \quad . \quad . \quad . \quad . \quad (12.8)$$

Of these, the only case of practical importance is the square shape

$$2b = h = \tfrac{1}{4}l \quad . \quad . \quad . \quad . \quad (12.9)$$

The author has endeavoured to compare these conclusions with regard to optimum shape with those arrived at by L. S. Palmer in his work on rectangular frame aerials.[*] Comparison is made difficult by the fact that Prof. Palmer's results are stated throughout in terms of current and the observations recorded are in terms of a current flowing, not in the loop itself, but in a kind of tuning circuit formed, apparently, by connecting a tuning condenser across a short length of the conductor of the loop. This branch circuit can be regarded as constituting a kind of tuning circuit, but it is not certain that its range of variation was sufficient for producing a true tuned condition over the whole range of frame dimensions examined.

In Fig. 6 of the paper referred to, a large number of current maxima are recorded, as functions of width for various heights, the wavelength being stated as 8·65 m. For one set of maxima the total lengths are approximately 15·6, 16, 16·4, 17·6, and 18 m. These are all in the neighbourhood of 2λ and these maxima are in accordance with paragraph 6 of Section (10). The maximum-maximum for l = approximately 2λ seems to occur for h just less than $\tfrac{1}{2}\lambda$, as compared with $\tfrac{1}{2}\lambda$ on the present theory. Other current maxima occur, however, with $l = 24\cdot8$, $24\cdot3$, $23\cdot4$, $24\cdot4$, which are in the neighbourhood of, though less than, $l = 3\lambda$. On the present theory the effective impedance is certainly a minimum under these conditions, but the effective e.m.f. is also very small, and current maxima would not be expected. The difference may arise from tilt of the field in Palmer's experiments. Minima are recorded for total lengths in the neighbourhood of λ and 5λ/2, which is also to be anticipated from the present theory. Thus some at least of the conclusions of the present theory are borne out by Prof. Palmer's observations, to a degree of accuracy as good as could be expected in view of the simplifying assumptions of the theory, namely that the effects of mutual interactions of the different parts of the aerial are adequately represented by the basic differential equations.

(13) ASYMMETRY AS A FUNCTION OF SIZE AND SHAPE

For most practical purposes it is desirable that the asymmetrical current i_g shall be as small as possible. One general method of achieving this is to separate the neutral point from earth by as high an impedance as possible, and to keep to a minimum the e.m.f.'s induced anywhere except in the frame aerial itself. It will, however, be of interest to examine the dependence of i_g on the size and shape of the frame aerial.

In the first place, that part of the asymmetrical e.m.f. which depends only on the frame and not on the neutral point connections to earth contains the term $\cos 2\pi b/\lambda$, and therefore vanishes whenever $2b = \tfrac{1}{2}\lambda$ or any integral multiple of this. It becomes small when $h = \lambda$, but this is not a practical case as the symmetrical e.m.f. is also small for this condition. It becomes large when $l = \lambda$, a condition which has already been discarded on other grounds.

A suggestion of some practical interest is the connection of the neutral point to earth by a half-wavelength or a wavelength line. In this case the e.m.f. due to the earth lead reduces to $\alpha\lambda^2/(4\pi)$ or $\alpha\lambda^2/(2\pi)$ and is therefore very small. At the same time the corresponding term in the impedance becomes very small, but if in addition $l = 2\lambda$, the term $\tfrac{1}{2}Z \coth \tfrac{1}{2}Pl$ in the impedance to i_g becomes very large. Thus the condition $2b = h = \tfrac{1}{2}\lambda$ and $g = \lambda$ or $\tfrac{1}{2}\lambda$ appears to have every advantage, since it minimizes both components of the unsymmetrical e.m.f. and opposes a high impedance to the residue.

(14) RECTANGULAR FRAME AERIAL: UNBALANCED TUNING

The equations for the currents i_0 and i_g for a frame as in Fig. 2, but with the left-hand tuning condenser replaced by a short-circuit, can be derived from (4.12) and (4.13) by putting $Z_3 = 0$, i.e. $\coth Pa_3 = 0$, and by summation of the e.m.f.'s acting on the vertical sides, as in Section (6). This process gives

$$i_0(Z_0 + Z \tanh Pl) = \frac{2_e}{P}\cdot\frac{\sinh(\tfrac{1}{2}Ph)}{\cosh Pl}$$
$$\left\{\epsilon^{-j\theta}\cosh\frac{3Pl}{4} - \epsilon^{j\theta}\cosh\frac{Pl}{4}\right\} + 2v_g\frac{\sinh^2(\tfrac{1}{2}Pl)}{\cosh Pl} \quad . \quad (14.1)$$

and

$$i_g(Z_0 + Z \tanh Pl) = \frac{2e}{P}\cdot\frac{\sinh\tfrac{1}{2}Ph}{\cosh Pl}$$
$$\left\{4\sinh(\tfrac{1}{2}Pl)\sinh(\tfrac{1}{4}Pl)\cos\theta\right\}$$
$$+ Z_0\left[\epsilon^{j\theta}\sinh(\tfrac{3}{4}Pl) - \epsilon^{-j\theta}\sinh(\tfrac{1}{4}Pl)\right]$$
$$+ v_g\left\{\frac{Z_0}{Z}\tanh Pl + 4\frac{\sinh^2(\tfrac{1}{2}Pl)}{\cosh Pl}\right\} \quad . \quad . \quad (14.2)$$

where $\theta = 2\pi b/\lambda$.

The formulae are complicated and the system appears to be, in general, ill adapted for either field-strength measurement or direction-finding, and it is not therefore proposed to analyse these formulae at any length. Some general observations may, however, be made.

So far as direction-finding is concerned, we have the well-known result that the received signal is not zero when the frame is parallel to the wave-front, i.e., putting $\theta = 0$,

$$i_0(Z_0 + Z \tanh Pl) = \frac{2e}{P}\,\frac{\sinh(\tfrac{1}{2}Ph)\sinh(\tfrac{1}{2}Pl)\sinh(\tfrac{1}{4}Pl)}{\cosh Pl}$$
$$+ 2v_g\frac{\sinh^2(\tfrac{1}{2}Pl)}{\cosh Pl} \quad . \quad . \quad . \quad (14.3)$$

This is the familiar "antenna effect." It may, however, be pointed out that the effect becomes very small when $l = 2\lambda$, a condition which also gives a large maximum value. Thus a frame with $2b = h = \tfrac{1}{2}\lambda$, earthed by a vertical line of length λ or $\frac{1}{2}\lambda$, might be worth considering for direction-finding on a fixed wavelength. In fact, such a frame has characteristics similar to those for the corresponding case with symmetrical tuning, which obviously arises from the fact that the effective impedance becomes very small in both cases and the required tuning impedance Z_0 correspondingly small.

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[*] See Reference (3).

(15) CIRCULAR LOOP: BALANCED TUNING

The formulae can be derived from (5.2) and (5.3) by a process of integration. Referring to Fig. 5, the field is assumed vertical and of zero phase at the vertical diameter. The tangential components of the fields at P and P′ will be e sin ϕ and $-e$ sin ϕ respectively, and the phases $-\theta$ and θ respectively, where $\theta = 2\pi r \sin\phi/\lambda$. From (5.2) and (5.3), the element di of i due to the summation for the two elements $rd\phi$ at P and P′ will be

$$di\,\frac{\cosh P(a_0+\frac{1}{2}l)}{\sinh Pa_0}$$

$$= -j\frac{e}{Z}r\sin\phi\cosh Pr(\pi-\phi)\sin\left(\frac{2\pi r}{\lambda}\sin\phi\right)d\phi \quad (15.01)$$

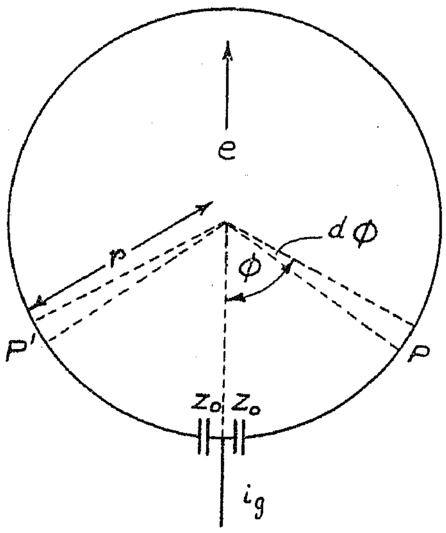

Fig. 5

Therefore

$$i\,\frac{\cosh P(a_0+\frac{1}{2}l)}{\sinh Pa_0}$$

$$= -j\frac{e}{Z}r\int_0^{\pi}\sin\phi\cosh Pr(\pi-\phi)\sin\left(\frac{2\pi r}{\lambda}\sin\phi\right)d\phi \quad . \quad (15.02)$$

For the general case when $2\pi r/\lambda$ is not an integer, it will be a close approximation to put

$$\cosh Pr(\pi-\phi) \simeq \cos\frac{2\pi r}{\lambda}(\pi-\phi) \quad . \quad (15.03)$$

i.e. $$i\,\frac{\cosh P(a_0+\frac{1}{2}l)}{\sinh Pa_0}$$

$$= -j\frac{e}{Z}r\int_0^{\pi}\sin\phi\cos\frac{2\pi r}{\lambda}(\pi-\phi)\sin\left(\frac{2\pi r}{\lambda}\sin\phi\right)d\phi$$

$$= -j\frac{e}{Z}r\int_0^{\pi}\sin\phi\cos\frac{2\pi r\phi}{\lambda}\sin\left(\frac{2\pi r}{\lambda}\sin\phi\right)d\phi \quad . \quad (15.04)$$

The integration can be carried out, for all cases in which $2\pi r/\lambda$ is not an integer, by means of the well-known formula

$$\sin(n\sin\phi) = 2\sum J_p(n)\sin p\phi;\ p = 1, 3, 5, \text{etc.} \quad . \quad (15.05)$$

and the result is

$$i\,\frac{\cosh PC\,a_0+\frac{1}{2}l}{\sinh Pa_0}$$

$$= j\frac{e}{Z}r\sum\frac{-4\,pn\sin n\pi}{\left\{(p+1)^2-n^2\right\}\left\{(p-1)^2-n^2\right\}};\ p = 1, 3, 5, \text{etc.} \quad . \quad . \quad (15.06)$$

where $$n = 2\pi r/\lambda \quad . \quad . \quad . \quad . \quad (15.07)$$

Thus the effective induced e.m.f. is given by

$$e_e = -8jnr\tan n\pi\,.\,e\,.\sum\frac{pJ_p(n)}{\left\{(p+1)^2-n^2\right\}\left\{(p-1)^2-n^2\right\}} \quad . \quad . \quad (15.08)$$

As a confirmation, it may be noted that for small values of n this tends to the usual value given in (1.1), i.e. the limit of the summation when $n \rightarrow 0$ is $-\frac{1}{8}n$.

Therefore $$|e_e| \rightarrow 8nrn\pi\frac{e}{8n} = n\pi re$$

$$= \frac{2\pi}{\lambda}Ae \quad . \quad . \quad . \quad (15.09)$$

where $A = \pi r^2 =$ area of loop.

If for any reason the more exact formula is required, i.e. a formula which includes the small effect of the resistance terms on the induced e.m.f., then since

$$\cosh Pr\phi = \cos(n - j\alpha r)\phi \quad . \quad . \quad (15.10)$$

the more exact formula can be derived by substituting $(n - j\alpha r)$ for n in the above, except in $J_p(n)$, which is not affected.

Certain special cases will be of interest. Deriving the formula for e_e from (15.02) (as in Section 7) gives

$$e_e = \frac{2jre}{\cosh Pr\pi}\int_0^{\pi}\cosh Pr\phi\sin\phi\sin(n\sin\phi)d\phi \quad (15.11)$$

$$\simeq \frac{2jre}{\cos n\pi}\int_0^{\pi}\cos n\phi\sin\phi\sin(n\sin\phi)d\phi \quad . \quad (15.12)$$

(1) $l = \lambda$; i.e. $n = 1$.

$$e_e = -2jre\cdot\frac{1}{2}\int_0^{\pi}\sin 2\phi\sin(\sin\phi)d\phi = 0 \quad (15.13)$$

In practice e_e will not be zero but will have a very small value depending on the resistance term α. It may be noted that for a rectangular loop with $l = \lambda$, e_e was also very small.

(2) $l = 2\lambda$; i.e. $n = 2$.

$$e_e = 2jre\int_0^{\pi}\cos 2\phi\sin\phi\sin(2\sin\phi)d\phi \quad (15.14)$$

$$= 2jre\frac{\pi}{2}\left\{J_3(2) - J_1(2)\right\} \quad . \quad . \quad . \quad . \quad (15.15)$$

$$= j \times 0{\cdot}448\lambda e \quad . \quad . \quad . \quad . \quad . \quad . \quad (15.16)$$

This may be compared with the formula already determined in Section (6) for a square coil with $l = 2\lambda$, i.e.

$$e_e = j\frac{2}{\pi}\lambda e = j \times 0\cdot 637\lambda e$$

Finally, in Section (10), it was shown that for a square frame with symmetrical tuning, the total length of conductor being $\frac{1}{4}\lambda$, the effective e.m.f. would be about 23 % greater than that calculated by the simple formula. By formula (15.08) the same ratio can be calculated for the circular loop of the same total length. It is found that for $n = \frac{1}{4}$ only the first term of the series is significant, and equals $0\cdot 504$ (i.e. approximately $\frac{1}{8}n$), and the e.m.f. in this case is about 28 % greater than that given by the simple formula. Thus changing the shape from square to circular only changes the effective e.m.f. by about 5 %.

(16) A NOTE ON THE MEASUREMENT OF THE OUTPUT VOLTAGE

It was shown in Section (6) that the currents i_0 and i_3 through the equal tuning impedances on either side of the mid or "earth" point on a symmetrically tuned frame aerial can be put in the form

$$i_0 = i + \tfrac{1}{2}i_g$$
$$i_3 = i - \tfrac{1}{2}i_g$$

The term i_g depends on the nature of the earth connection, if any, and in any case on the impedance to earth of any apparatus connected to the mid-point. Thus, for field-strength measurement, unless the current i_g can be made negligibly small it is in general necessary to be able to measure the vector sum of the unequal voltages produced across the two halves of the tuning impedance in series.

Various means of doing this suggest themselves. For example, the voltages can be applied to the grids of a push-bell amplifying circuit of which the output is a single voltage proportional to the vector sum of the input voltages. Such an arrangement, however, though easy to design in principle, is not easy to realize in practice, particularly to the shorter wavelengths to which most of the considerations in this analysis apply. Further, it must be pointed out that the direct application of the voltages to the terminals of a "push-pull" rectifier circuit will not in general give an output proportional to the vector sum of the input voltages.

It can, however, be shown on theoretical grounds that a push-pull frequency-change system may satisfy the necessary condition, provided that a close approximation to electrical symmetry in the two halves of the system can be realized. Similarly the multi-grid frequency-change valves in which the frequency-change mechanism is essentially the multiplication of a signal voltage and a local oscillator voltage, will, under the same assumed conditions, give an output proportional to the vector sum of the input voltages. In both cases, however, considerable care will be necessary to realize the assumed conditions.

(17) CONCLUSIONS

On the assumption that the current and voltage relationships for a single conductor bent into a closed shape—rectangle or circle—can be represented by the partial differential equations of transmission-line theory, formulae have been developed for the effective induced e.m.f. and effective aerial impedance of such closed aerials, without limitation of linear dimension. Subject to experimental confirmation, the work is considered to serve two main objects. (1) It should enable some estimate to be made of the size at which a frame aerial, as used for field-set strength measurement, ceases to be "small" in the generally accepted sense, and it should also enable the e.m.f. induced in such a "small" frame to be calculated, in terms of the field intensity, somewhat more accurately than by the usual simple formula. (2) It should serve as a preliminary guide in deciding on the type of aerial best suited to field-strength measurement at short and ultra-short wavelengths.

In the only existing internationally agreed recommendation on the latter subject (Commission I, U.R.S.I., 1934), the loop aerial is made the standard "collector" for field-strength measurement for wavelengths longer than 15 m., and "a di-pole capable of orientation" for all other wavelengths. It may be, however, that a loop aerial, of dimensions not necessarily small compared with the wavelength, would prove to have advantages. This is essentially a matter for practical experience, and the present theoretical analysis may serve as a guide for such experiments.

Certain other points have emerged from the analysis. It has been shown, for example, that a loop receiving system having complete geometrical symmetry will not necessarily show electrical symmetry in its "maximum" orientation even in a uniform wave-field; i.e. the currents and voltages will not, in general, be symmetrical about the vertical axis of geometric symmetry. This is true in general even for small loop aerials, but will not in general be of practical significance in such cases. It is, however, a feature which calls for some care in short-wave and ultra-short-wave intensity measurements.

This asymmetry is of course already familiar in the use of coil aerials for direction-finding. The formulae give the magnitude of the "antenna effect" in any given case and confirm, moreover, that the establishment of geometrical and electric-circuit symmetry should suffice to eliminate the effect in the "minimum" orientation of the loop aerial.

(18) ACKNOWLEDGMENTS

The author is indebted to Dr. R. L. Smith-Rose for valuable comment and suggestion. The work was carried out as part of the programme of the Radio Research Board and is published by permission of the Department of Scientific and Industrial Research.

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THE DISTRIBUTION OF ULTRA-HIGH-FREQUENCY CURRENTS IN LONG TRANSMITTING AND RECEIVING ANTENNAE[*]

By Professor L. S. PALMER, D.Sc., Ph.D., Member, and K. G. GILLARD, B.Sc.

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SUMMARY

The distribution of ultra-high-frequency currents in long transmitting and receiving antennae is considered analytically, and experiments are described which were undertaken to test the conclusions arising from the theoretical analyses.

With an isolated transmitting antenna energized at any point, the current is found to be distributed sinusoidally in the two parts of the antenna which are separated by the point of energy input. The currents in the two parts are dependent upon each other in a manner determined by the approximate equations (3) and (4) in the paper.

With an isolated receiving antenna energized by a uniform electromagnetic field, it is found that the current distribution is such that current nodes occur at intervals of one wavelength measured from both ends of the aerial. The distribution is given approximately by equation (5) in the paper.

" End effects " and modifications produced by the experimental methods used are also discussed.

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(1) INTRODUCTION

In order to determine the radiating and receiving properties of a long antenna, it is necessary to know the current distribution along the antenna wire when it is used as a transmitting aerial and as a receiving aerial, respectively. This, in turn, presupposes a knowledge of the variation along the antenna of its impedance per unit length: a complex quantity which is not susceptible to any simple mathematical treatment.

Any exact theory must take into account such interrelated factors as the difference between the wave velocity along the antenna wire and in the surrounding medium due to the distributed electrical properties of the particular wire, the phase change on reflection of the waves due to localized electrical properties near the ends of the wire, the so-called " shortening factor " due, according to some authorities, to both distributed and localized effects, the possible reflection of the waves from the earth's surface, and other similar items all of which tend to make any exact theory extremely complicated. Various authors have considered one or more of these factors and, by suitable approximations, have evolved theories which, in the few cases where they have been tested, conform approximately to experimental data. When, however, experimental measurements at very high frequencies are attempted, the inherent difficulties arising from the use of ultra-short waves with long antennae introduce errors which are much greater than those arising from the usual theoretical approximations. For example, the introduction into an antenna of a current-measuring instrument produces effects at very high frequencies which considerably alter the current distribution anticipated by any of the published theories. In fact, it is not possible, in many cases, to decide experimentally between, for example, the simple theory of the current distribution in earthed receiving antennae with uniform electrical constants put forward by Moullin[†] in 1925, or, say, the more elaborate theories of Metzler[‡] and Hara,[§] the former of whom in 1936 used transmission-line equations and resolved his resultant current into progressive and stationary waves, and the latter of whom (in the same year) calculated the radiation impedance of linear conductors assuming a sinusoidal current distribution. Again, Colebrook[||] in 1932 used transmission-line equations and gave three different values to the antenna electrical constants per unit length for three different parts of the antenna, and then considered special cases of his general formula; whilst Labus,[¶] in the following year, determined the aerial impedance from the line integral $\int E(x)I(x)dx$, where $I(x)$ is the current produced by the x component of the incident field E acting on an aerial with uniformly distributed electrical constants.

In no instance have the theoretical conclusions concerning current distributions in long transmitting or receiving aerials been tested experimentally at high frequencies, although in 1928 Wilmotte[**] measured the current distribution in the particular case of an earthed transmitting aerial which was 11 m. long and radiated waves as short as 15·6 m.

It was therefore felt desirable to attempt to reduce experimental errors as much as possible, and then to employ the simplest theory consistent with the fact that the errors arising from the theoretical assumptions should not exceed those arising from the experimental methods adopted. The following simple theoretical treatment complied with this limitation, and in Section (4) the results of experiments on wavelengths of about 70 cm. are compared with the theoretical deductions.

With greater precision in the future it is hoped to be able to verify experimentally some of the more rigorous theoretical expressions which have been deduced by other authors, but this was not possible with the methods and apparatus at present available.

(2) THEORETICAL CONSIDERATIONS

The essential distinction between long transmitting and long receiving antennae when used with ultra-short waves depends on the fact that in the former case the input e.m.f. is concentrated over a short length of

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[*] Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 415.

[†] See Reference (1).

[‡] *Ibid.*, (2).

[§] *Ibid.*, (3).

[||] *Ibid.*, (4).

[¶] *Ibid.*, (5).

[**] *Ibid.*, (6).

the antenna wire, whilst in the latter case it is usually uniformly distributed throughout the whole length of the wire. These two particular cases will now be considered with a view to determining the approximate current distributions which result.

(A) The Long Transmitting Antenna

Consider an isolated antenna AB (Fig. 1) of length greater than one wavelength, and let it be energized at any point P which is at a distance D from the lower end B.

As it is intended to simplify the theoretical treatment as much as possible owing to the unavoidably large errors inherent in ultra-high-frequency measurements, it will be assumed that the electric constants of the antenna wire are uniform throughout its length and that the resistance and leakance are negligible compared with the inductive and capacitive reactances respectively.

Then if L and C denote the inductance and capacitance per unit length of the antenna wire, the fundamental equations for the voltage V and current I at any arbitrary point Q are given by

$$-dV = j\omega LIdx \quad \text{and} \quad -dI = j\omega CVdx$$

where $j = \sqrt{(-1)}$ and ω is 2π times the frequency.

Thus $d^2V/dx^2 = -\omega^2 LCV$ and $d^2I/dx^2 = -\omega^2 LCI$.

Hence $I = S\cos 2\pi x/\lambda + T\sin 2\pi x/\lambda$

where λ is the wavelength and S and T are the integration constants. These can be determined from the particular boundary conditions at A, B, and P. Let the current in the portion AP be I_A and that in the portion PB be I_B. Then these conditions are:—

(i) At A, $x = l$ and $I_A = 0$.
(ii) At B, $x = 0$ and $I_B = 0$.
(iii) At P, $x = D$ and $I_A = I_B = I_0 e^{j\omega t}$ (say).
(iv) At P, $x = D$ and $z_0\Delta(dI/dx) = E_0 e^{j\omega t}$,

where z_0 is some constant. This follows from the fact that any change in the current gradient at P must be compensated by a voltage input $E_0 e^{j\omega t}$. Thus the fourth boundary condition may be written

$$E_0 e^{j\omega t} = z_0\left(\frac{dI_A}{dx} - \frac{dI_B}{dx}\right)$$

It will be necessary, in view of this current change at P, to treat separately the current in those parts of the antenna above and below P.

From boundary conditions (ii) and (iii), we have

$$S = 0 \quad \text{and} \quad T = I_0 e^{j\omega t}/\sin 2\pi D/\lambda$$

Hence
$$I_A = \frac{\sin 2\pi(l-x)/\lambda}{\sin 2\pi(l-D)/\lambda} I_0 e^{j\omega t} \quad . \quad . \quad . \quad (1)$$

and
$$I_B = \frac{\sin 2\pi x/\lambda}{\sin 2\pi D/\lambda} I_0 e^{j\omega t} \quad . \quad . \quad . \quad . \quad (2)$$

That is, if the boundary conditions (i), (ii), and (iii), are alone considered, the currents I_A and I_B should vary sinusoidally and independently along the two parts of the antenna with zero current at the open ends remote from the input e.m.f. at P. The present problem is, however, more complicated because the e.m.f. at P, unlike an e.m.f. localized at the foot of an aerial, does not alone prescribe the motions of the charges in each portion of the antenna. For example, the current I_A is not only dependent on the e.m.f. at P but is also affected by the current I_B in the portion of the antenna PB. As a consequence of this we must employ the boundary condition (iv), which gives the relation between the change of current gradient at P and the input e.m.f., and so determines a relation between E_0 and I_0.

On differentiating equations (1) and (2) with respect to x, and using condition (iv), we get the required relation between E_0 and I_0, namely

$$I_0 = \frac{\lambda E_0}{2\pi z_0} \cdot \frac{(\sin 2\pi D/\lambda)\sin 2\pi(l-D)/\lambda}{\sin 2\pi l/\lambda}$$

Hence

$$I_A = \frac{\lambda}{2\pi} \cdot \frac{E_0}{z_0}\left[\frac{\sin 2\pi D/\lambda}{\sin 2\pi l/\lambda} \sin \frac{2\pi(l-x)}{\lambda}\right] \sin(\omega t + \pi) \quad (3)$$

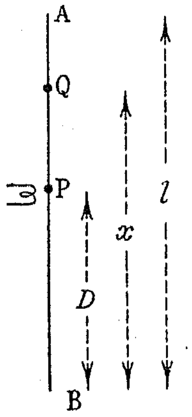

Fig. 1

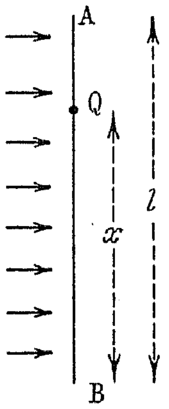

Fig. 2

and

$$I_B = \frac{\lambda}{2\pi} \cdot \frac{E_0}{z_0}\left[\frac{\sin 2\pi(l-D)/\lambda}{\sin 2\pi l/\lambda} \sin \frac{2\pi x}{\lambda}\right] \sin(\omega t + \pi) \quad (4)$$

From these two equations three conclusions immediately follow:—

(a) Firstly, the current in each portion of the antenna varies sinusoidally and is zero at the open ends remote from the e.m.f. at P. This conclusion has already been deduced and tested by Wilmotte* for the particular case when P coincides with the point B (Fig. 1), and by Metzler† when P is mid-way between A and B.

(b) Secondly, it follows from equations (3) and (4) that the two currents I_A and I_B are not independent of each other. If the length of one portion of the antenna be an integral number of half-wavelengths (say AP $= n\lambda/2$), then the current I_B in the other portion PB becomes zero for all values of x, whatever its length (except in the special case where PB is also an integral number of half-waves in length). Similarly, when PB (or D) $= n\lambda/2$, $I_A = 0$ throughout the length AP (except when AP also equals $n\lambda/2$).

Owing to the fact that in any given antenna the phase-change on reflection from the open ends of the aerial wire will not be exactly π, the current will not be zero when $x = 0$ or l, and consequently the current I_A, for example, will attain a minimum value for lengths PB or D which will be somewhat less than $n\lambda/2$. Furthermore the finite values of the antenna resistance and leakance will affect the magnitude of the currents in such a way that the maximum values will be reduced and the minimum values will not be zero.

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* See Reference (6).

† *Ibid.*, (2).

Again, when the length of one portion (PB say) of the antenna is an odd number of $\frac{1}{4}$-wavelengths, the current I_A in the other part will be a maximum. That is, if $D = (2n + 1)\lambda/4$, I_A will be a maximum, or if $(l - D) = (2n + 1)\lambda/4$, I_B will be a maximum. As before, these critical values will be slightly modified because the phase-change on reflection is not exactly π, and the antenna resistance and leakance are not quite negligible.

Metzler* has treated this problem of the long transmitting antenna from a somewhat different standpoint and has studied the effects of resistance and leakance by analysing his expression for the antenna current into progressive and stationary waves; the former determine the slight modifications produced on the simple stationary-wave system due to currents passing through the nodal points in order to compensate for resistance and leakance losses in the antenna.

(*c*) A third conclusion follows if in equation (3) I is a function of l, and D and $(l - x)$ are kept constant. Then $I_A \propto \operatorname{cosec} 2\pi l/\lambda$. That is, if the point Q at which I_A is measured is at a constant distance from the end A, and the point P where the e.m.f. is applied is at a constant distance from the end B, then the current measured at Q should be proportional to $\operatorname{cosec} 2\pi l/\lambda$.

Thus by three different experimental methods based on the above three conclusions it should be possible to check the foregoing theoretical equations (3) and (4). Such experiments are described in Section (3), and the results are discussed in Section (4).

(B) The Long Receiving Antenna

Consider a long isolated receiving antenna AB energized throughout its length by a uniform e.m.f. $E_0 e^{j\omega t}$ per unit length. With the same assumptions and nomenclature as before, the fundamental equations for the voltage V and current I at any arbitrary point Q will now be

$$-dV = j\omega LIdx - E_0 e^{j\omega t}dx$$

and

$$-dI = j\omega CVdx$$

Thus

$$\frac{d^2V}{dx^2} = -\omega^2 LCV \text{ as before, but}$$

$$\frac{d^2I}{dx^2} = -\omega^2 LCI - j\omega CE_0 e^{j\omega t}$$

Hence

$$V = S' \cos 2\pi x/\lambda + T' \sin 2\pi x/\lambda$$

and

$$I = \frac{E_0}{\omega L} e^{j\omega t} + S'\sqrt{\left(\frac{C}{L}\right)} \sin 2\pi x/\lambda + T'\sqrt{\left(\frac{C}{L}\right)} \cos 2\pi x/\lambda$$

S' and T' are the integration constants to be determined from the particular boundary conditions, which are, in this case,

(i) at A: $x = l$ and $I = 0$.
(ii) at B: $x = 0$ and $I = 0$.

Hence from (ii) $$T' = -\frac{\lambda E_0}{2\pi} e^{j\omega t}$$

and from (i) $$S' = -\frac{\lambda E_0}{2\pi} e^{j\omega t} . \tan \pi l/\lambda$$

Thus

$$I = \frac{2E_0}{\omega L} \sin \frac{\pi x}{\lambda}\left(\sin \frac{\pi x}{\lambda} - \tan \frac{\pi l}{\lambda} \cos \frac{\pi x}{\lambda}\right) \cos \omega t$$

$$= \frac{2E_0}{\omega L}\left[\sec \frac{\pi l}{\lambda} \sin \frac{\pi x}{\lambda} \sin \frac{\pi(l - x)}{\lambda}\right] \cos (\omega t + \pi) \quad . \quad (5)$$

From this equation it follows that the current is not distributed sinusoidally with current nodes every $\frac{1}{2}$-wavelength, but that current nodes occur at intervals of a whole wavelength measured from both ends of the antenna. Looked at geometrically rather than physically, the current distribution may be considered to be represented by a sine wave with its axis parallel to but displaced from the antenna wire. This fact was first pointed out by Hagen and recorded by Korshenewsky.* Colebrook refers to it in the last section of a recent paper in the *Journal*.† The resulting theoretical current distributions for different values of l/λ are shown by the seven dotted graphs of Fig. 9.

The result given in equation (5) for the current at any point Q (Fig. 2) may also be obtained by integrating over the length l of the antenna the effect at Q of a current $I_0 dl$ at any other point P, with the assumptions that waves travel from P with the velocity $1/\sqrt{(LC)}$ in both directions and are reflected to and fro an infinite number of times along the antenna without loss of amplitude and with phase-change of π at the open ends of the antenna.

It was found that the simple theoretical result given by equation (5) could not be verified by direct experiment, firstly because the currents were not zero at the open ends of the antenna, or in other words, the phase-change on reflection of the waves from the ends of the antenna differed from π; and secondly, because the introduction of the current-measuring instrument appreciably modified the original current distribution. It is a comparatively simple matter to modify equation (5) by taking into consideration the fact that the phase-change on reflection is not exactly π. The error is that which would be produced if the aerial were extended by an amount $a/2$, say, at each end. The phase-change at the actual termination of the wire would be $2\left(\frac{2\pi}{\lambda} \cdot \frac{\alpha}{2}\right)$ or $2\pi a/\lambda$, and the appropriate boundary conditions are that the current is zero for values of x equal to $-\frac{1}{2}a$ and $(l + \frac{1}{2}a)$. With these conditions the current amplitude at any point Q is given by

$$I_0 = K\left[1 - \sec \frac{\pi(l + a)}{\lambda} \cos \frac{\pi a}{\lambda} \cos \frac{\pi(l - 2x)}{\lambda}\right] \quad (6)$$

where K is a constant. This expression reduces to the amplitude of the current given by equation (5) when $a \to 0$.

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* See Reference (2).

* See Reference (7). † *Ibid.*, (4).

Although equation (6) should be more exact than equation (5) for the current distribution, it is not possible to check it practically because of the instrumental error referred to above.

It was found experimentally that the introduction of a vacuum thermojunction into the antenna shortened its effective length by an amount $2b$, say. That is, the thermojunction acted as a length $2b$ of antenna wire through which the current circulated but along which the e.m.f. of the incident field was inoperative. Consequently, for the ends of this section of the antenna the appropriate boundary conditions are the continuity of current and current gradient and the discontinuity of e.m.f. If the thermojunction be inserted at the point Q (Fig. 2) and the "equivalent length" of the antenna be l' $(= l + 2b)$, then the necessary 12 boundary conditions are (at distances x' along the antenna):—

At B, where $x' = -\frac{1}{2}a$, $I_1 = 0$; $E_1 = E$.

$$\text{At Q}\begin{cases}\text{where } x' = (x - b),\ I_1 = I_2,\ I_1' = I_2';\ E_1 = E, \\ \quad E_2 = 0. \\ \text{where } x' = (x + b),\ I_2 = I_3,\ I_2', = I_3';\ E_2 = 0, \\ \quad E_3 = E.\end{cases}$$

At A, where $x' = (l + \frac{1}{2}a)$, $I_3 = 0$; $E_3 = E$.

The currents will differ in phase by the appropriate values of $2\pi x'/\lambda$, but will have constant amplitude, the resistance and leakage being neglected.

The resulting equation for the amplitude of the current at Q is

$$\begin{aligned} I_0 = K \operatorname{cosec} 2\pi(l' + a)/\lambda\big[&\sin 2\pi(l' + a - b)/\lambda \\ &+ \sin 2\pi b/\lambda \,.\, \cos 2\pi(l' - 2x)/\lambda \\ &- 2 \cos \pi a/\lambda \,.\, \sin \pi(l' + a)/\lambda \,.\, \cos \pi(l' - 2x)/\lambda\big] \quad (7)\end{aligned}$$

The expression on the right-hand side reduces to the amplitude in equation (5) when $a = b = 0$ and $l' = l$.

Equation (7) may also be deduced by the alternative method of summing the infinite series of waves which may be considered to travel to and fro along the antenna after reflection at the ends.

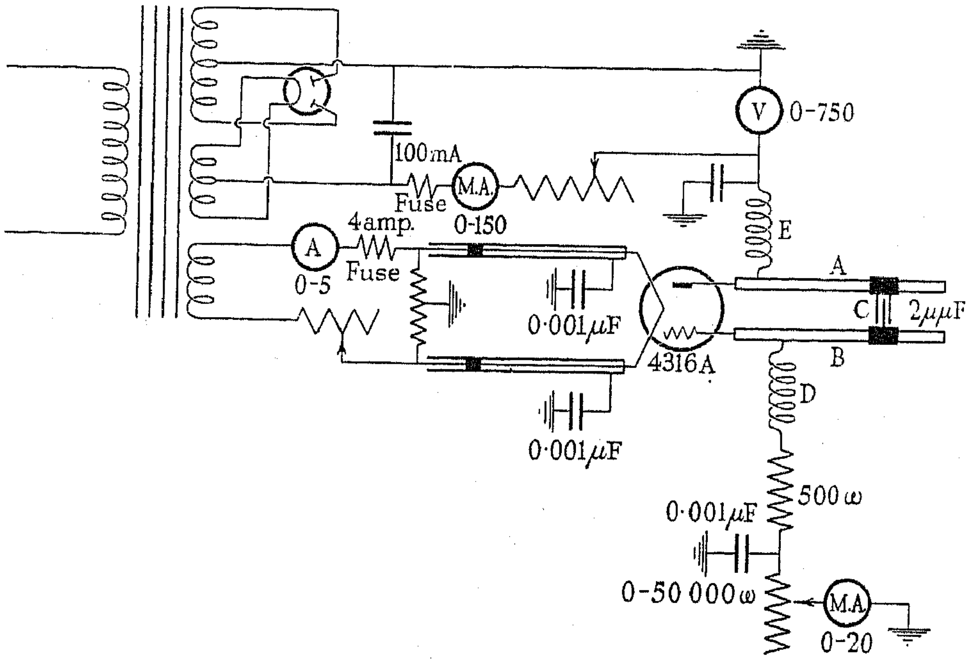

Fig. 3

The graphical representations of equation (7) are given by the seven full-line curves in Fig. 9, and may there be compared with the dotted graphs of equation (5).

(3) EXPERIMENTAL WORK

In order to test the foregoing theoretical conclusions it was decided to work with as short waves as possible so as to keep the dimensions of the apparatus as small as possible. Furthermore, the antennae could then be elevated several wavelengths above the earth's surface without any serious experimental difficulty.

A valve oscillator was constructed to work on wavelengths of the order of 60 to 100 cm. When testing the validity of equations (3) and (4), the oscillator was coupled indirectly to a straight brass-rod antenna of variable length, and when used in connection with the measurement of current distributions in receiving antennae [equations (5), (6), and (7)] it was coupled to an ordinary half-wave radiator with a cylindrical parabolic reflector. With these experiments the brass rod was used as the receiving antenna.

(A) The Oscillator

A G.E.C. 4316A valve was used with the modified Hartley circuit shown in Fig. 3. To the plate and grid were fastened two straight rods A and B, which were connected together by a small fixed tubular condenser C.

The length of the radiated wave could be varied by sliding this condenser along the rods, thereby changing the effective inductance in the circuit. The grid was connected to the filament by a variable resistance and a small choke D. The plate potential of 400–500 volts was transformed from the mains, rectified, and applied through the choke E to the junction of the inductance rod A and the plate lead. It was found that the filament chokes needed to be critically adjusted for each wavelength. Each filament lead was therefore constructed from a brass tube along the axis of which was a copper rod connected to the tube at one end and insulated from it at the other end. Along this rod and in contact with the inside of the tube, a small brass cylinder could slide. The low-tension supply was connected to the rod, and the filament lead to the outside of the brass tube. The general arrangement of these filament chokes is shown in Fig. 3. The radiating aerial was coupled to the oscillating circuit by placing it a centimetre or so from the tubular condenser C and parallel to it. The oscillator was calibrated by coupling it to Lecher wires along which could slide a circular reflecting copper screen. This altered the effective length of the wires and gave very sharp current peaks in a vacuum thermojunction which was inserted in a wire bridge also capable of sliding along the Lecher wires.

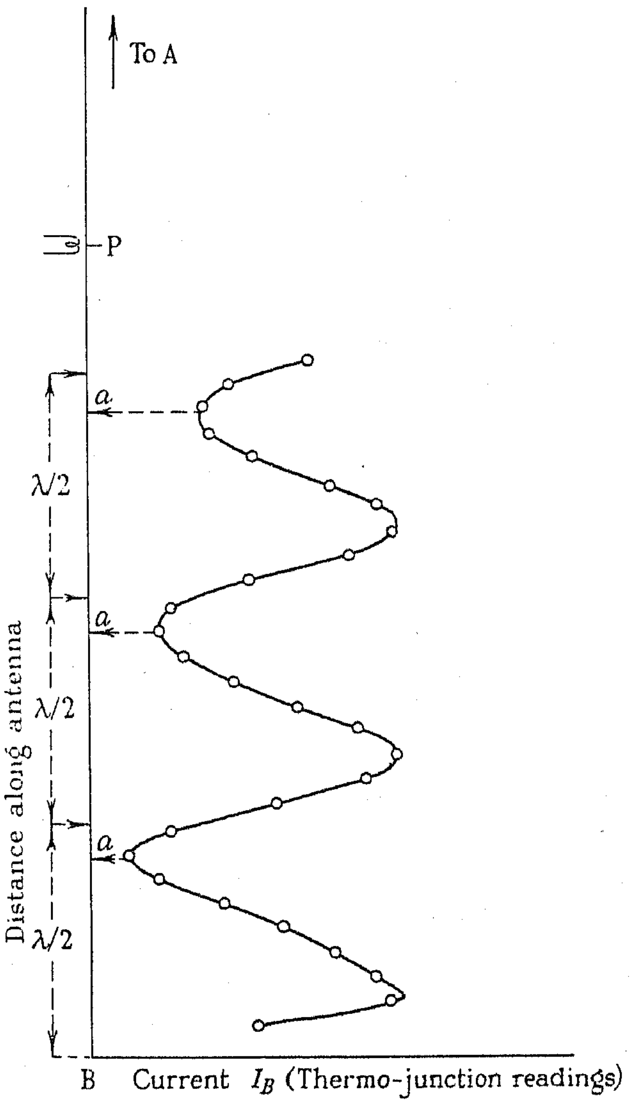

Fig. 4

$\lambda =$ 74 cm.
$PB =$ 130 cm.

(B) The Receiving Antenna

The brass-rod antenna was made in sections of varying lengths which screwed together in such a way that a vacuum thermojunction could be incorporated in the antenna at any required point, and could be moved along the antenna by unscrewing a portion from one end and screwing it on the other. The rod was 8 mm. in diameter. As an alternative method, the vacuum thermojunction was placed in a small bridge the ends of which slid along the antenna and made contact with it. The distance between the ends was about 2 cm.

(C) Experimental Procedure

With the transmitting antenna, experiments were made to test the three conclusions deduced from equations (3) and (4), namely:

(i) The sinusoidal nature of the current distribution, assuming $I = f(x)$.

(ii) The condition for maximum and minimum values of the currents I_A and I_B, assuming $I_A = f(PB)$ and $I_B = f(PA)$ respectively (see Fig. 1).

(iii) The dependence of the current on the total length of the antenna, assuming $I = f(l)$.

Finally, with the receiving antenna, the current distributions were measured for 7 values of l'/λ, varying from 0·86 to 2·51.

The results of the experiments with the transmitting antenna are shown in Figs. 4 to 8, and are discussed with reference to equations (3) and (4). The circles in Fig. 9 represent the experimental values of the current in a receiving antenna. Their distribution may be compared with the graphical representations of equations (5) and (7)—the dotted and full-line graphs respectively.

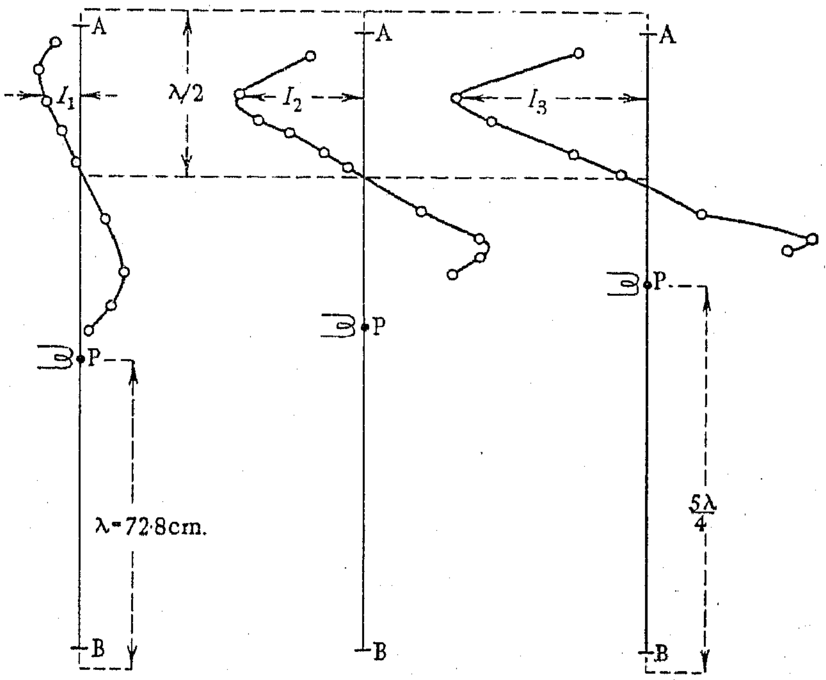

Fig. 5

(4) RESULTS

(A) Transmitting Aerial

(*a*). Fig. 4 shows the sinusoidal form of the current along one part of a transmitting antenna, and, except for the slight displacement *a* of the current curve, it confirms the accepted theoretical distribution for a distance of nearly 2 wavelengths. The displacement *a* is discussed when referring to Figs. 6 and 7 below. This graph was obtained by gradually moving a vacuum thermojunction along the antenna from B towards P, keeping the lengths AP and PB constant. The spacing of the current nodes at intervals of $\frac{1}{2}$ wavelength measured from the free end has already been shown by Wilmotte* for the first node in the case of an earthed transmitting antenna.

(*b*). Fig. 5 shows the same current distribution and also the fact that, in the portion PA of the antenna, the current I_A increases from a minimum I_1 to a maximum I_3 as the length of the other part PB increases from $n\lambda/2$ to $(2n + 1)\lambda/4$, n being 2 in this particular case. That the minimum value of I_1 is dependent on the length of the portion PB and is practically independent of the length PA in which the current I_1 is flowing, was shown by the fact that its variations (when PB is constant and equal to λ) were almost negligible as AP was increased in length from about 35 cm. to 110 cm. This is not because the portion AP has maximum impedance whatever be its length, but rather because the other portion PB has a constant minimum impedance, and the two portions may be considered to act as parallel loads. These variations of the current I_A in AP are more clearly shown in Fig. 6, where the length PB (or D) was gradually increased from about 8 cm. or less than $\frac{1}{8}$ wavelength to over 2 wavelengths. Currents I_1, I_2, and I_3, in this figure correspond to the values I_1, I_2, and I_3, respectively, in Fig. 5. The

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* See Reference (6).

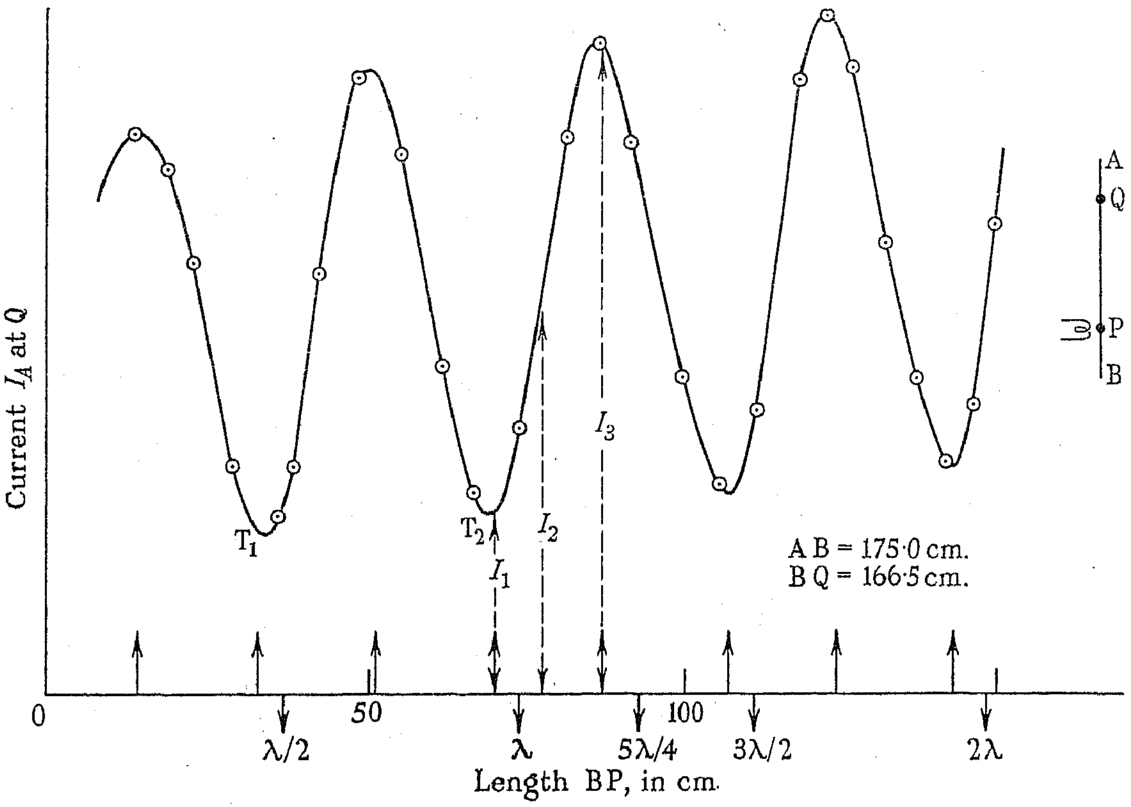

Fig. 6

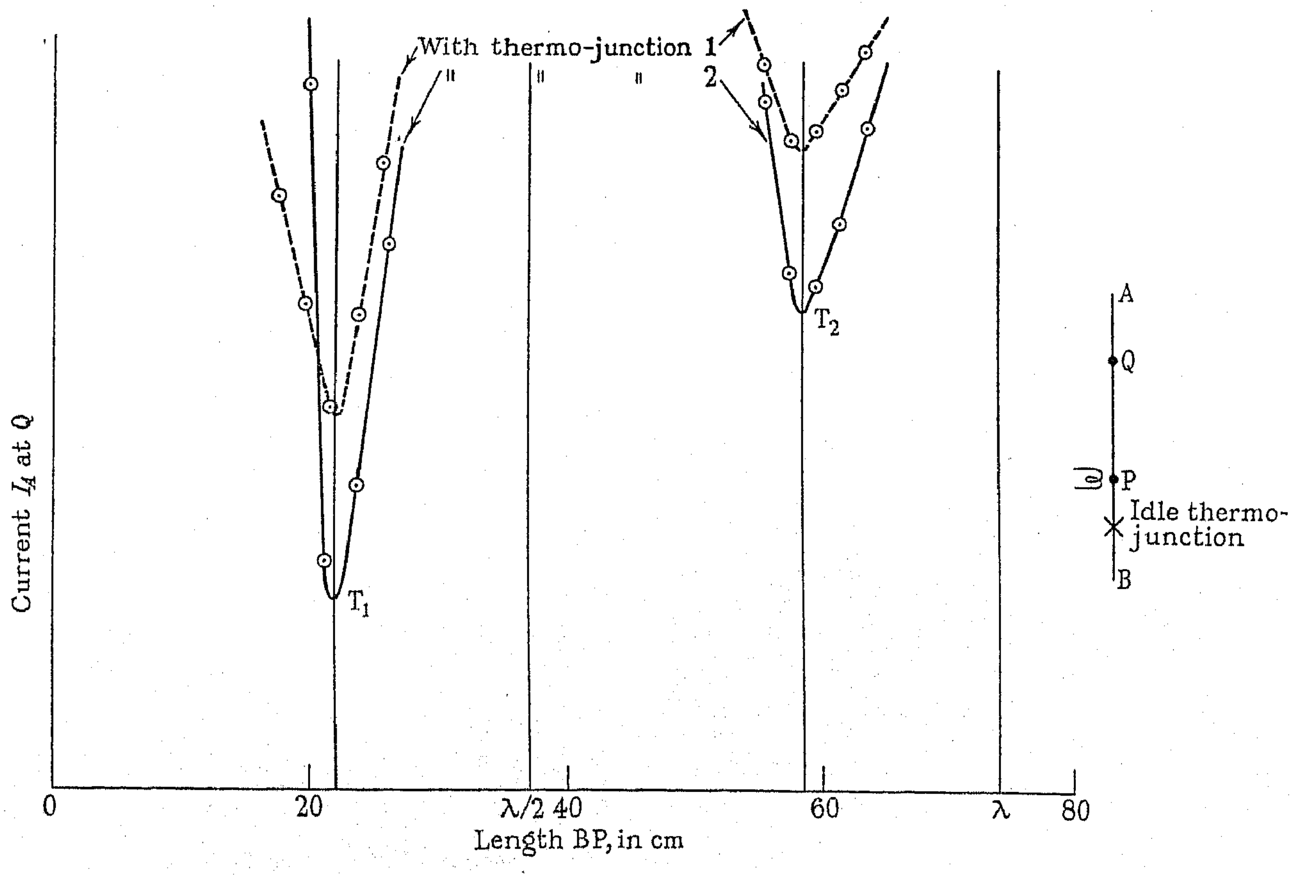

Fig. 7

AB = 175 cm.
BQ = 166·5 cm.

current I_A varies periodically with distance, the period being $\frac{1}{2}$ wavelength, but the turning points are displaced by 4 cm. from the positions calculated from the wavelength as measured on Lecher wires. The fact that the "equivalent" length of an antenna is greater than the geometrical length has been ascribed to various causes. Tani* states that "the wavelength of the wave on the antenna wire is markedly shortened as antenna insulators of large capacitance are used." Metzler† uses a reduced wavelength scale dependent on the velocity of wave propagation along the wires, whilst Hara‡ ascribes the effect to reflection from the earth and to the thickness of the aerial wire. All these factors produce a kind of shortening effect which is operative throughout the length of the antenna and may be measured in equivalent centimetres of antenna wire. But with the present experiments such effects are negligible compared with that produced at the ends of the antenna, and by far the largest change in the equivalent length was produced by altering the shape of the end of the wire. We believe that over 90 % of the effect referred to here is due to the fact that, on reflection of the waves from the end of the antenna, the phase-change is not exactly π and the resultant current at the end is not zero. Tani referred to a similar end-effect and stated that he obtained a shortening factor of 3·8 % when "a capacitance of 2·87 $\mu\mu$F was placed at the end of the antenna." It would seem desirable to distinguish the distributed "shortening effect" from the localized "end effect." It is significant that, when both ends of the antenna are involved, the effect is almost exactly doubled. The value of 4 cm. is therefore the value of a in equation (6) and the equivalent phase-change on reflection with this antenna was therefore $1\cdot 11\,\pi$, the wavelength being 72·8 cm. Apart from this "end effect," which is now being investigated further, Figs. 4, 5, and 6, give support to the conclusions (*a*) and (*b*) deduced on page 286 from equations (3) and (4).

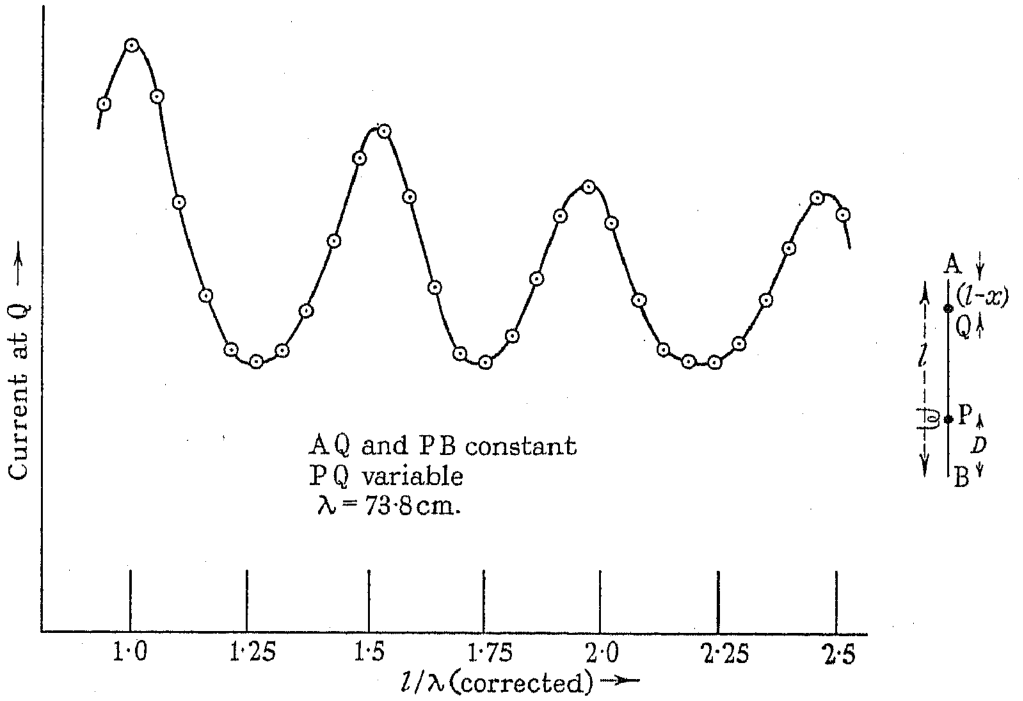

Fig. 8

Fig. 7 is similar to a part of Fig. 6, and shows the new positions of troughs T_1 and T_2 in the latter figure when an idle thermojunction (with leads and ammeter) was inserted in the portion PB of the antenna. The difference in the positions of these troughs in Figs. 6 and 7 gives the equivalent length of the idle thermojunction in centimetres of antenna wire. From the figure this is seen to be about 11 cm. for each junction. Hence b in equation (7) is equal to 5·5 cm.

Thus a total correction of 19 cm. is necessary when both corrections arise, as, for example, in the current distribution measurements in a long receiving antenna where there are two end-effects and one thermojunction correction.

(*c*). The third deduction which was made from equations (3) and (4) was based on the fact that $I_A \propto \operatorname{cosec} 2\pi l/\lambda$ when D and $(l-x)$ in equation (3) were kept constant. This was tested experimentally by gradually increasing the length of the portion PQ of the antenna (Fig. 1), thereby keeping AQ [or $(l-x)$] and BP (or D) constant. The current should reach infinity for values of $l/\lambda = n\lambda/2$ and should have some constant minimum values for $l/\lambda = (2n+1)\lambda/4$. Resistances will of course modify these theoretical conclusions concerning infinite current values, and the experimental results (corrected for the end-effects and thermojunction) are shown in Fig. 8.

(B) Receiving Aerial

For testing equations (5) and (7) a vacuum thermojunction was gradually moved along the rod antenna which was used as a receiving aerial at a distance of several wavelengths from the transmitter. To move the thermojunction a portion of the rod was unscrewed from one end and screwed in the other, and the aerial adjusted so that its position remained fixed with respect to the transmitter.

In Fig. 9 the experimental values so obtained are

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* See Reference (8). † *Ibid.*, (2). ‡ *Ibid.*, (3).

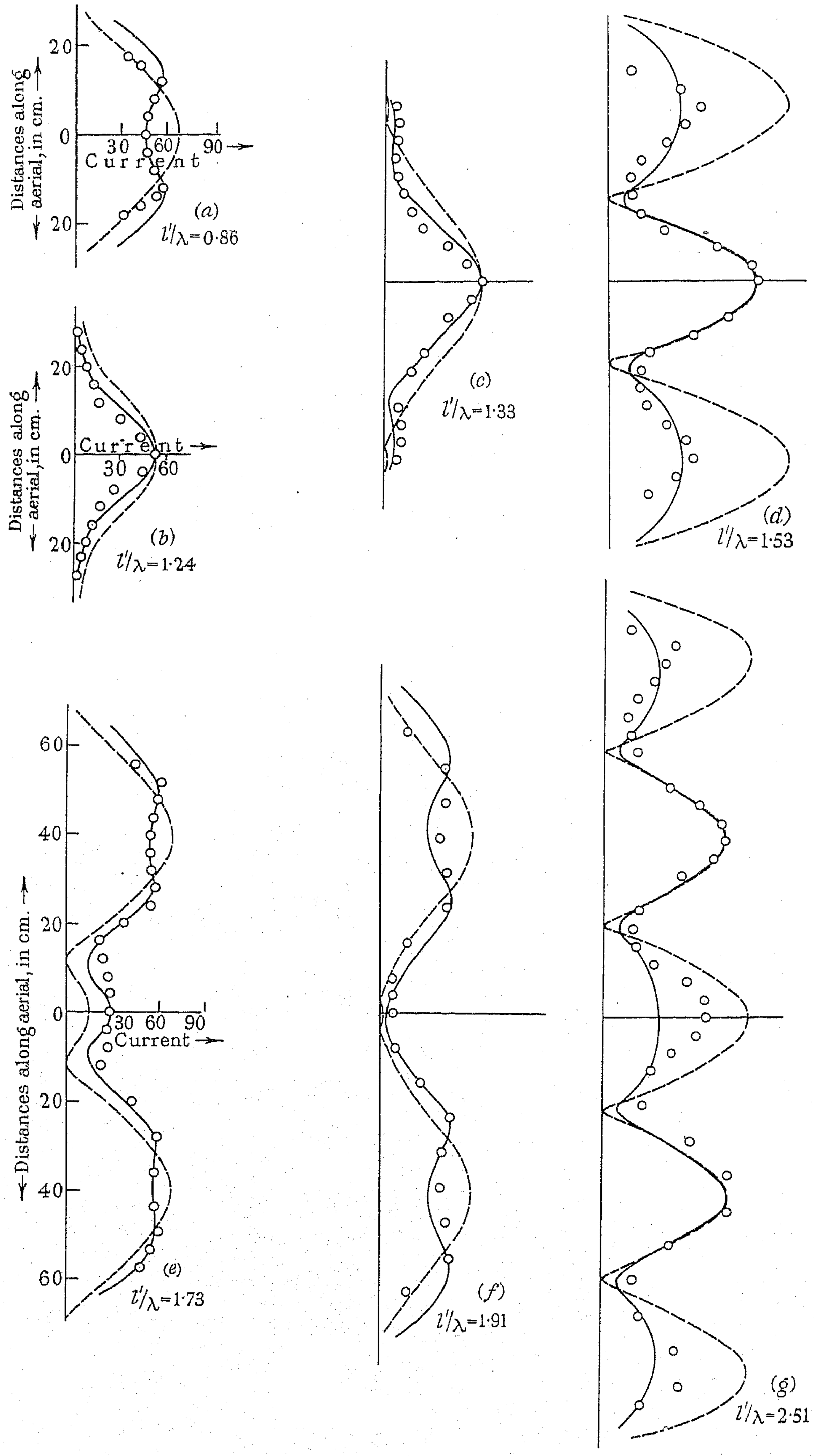

Fig. 9

- - - equation (5).
—— equation (7).
○○○ experimental measurements.

shown by the small circles, whilst the graphs are plotted from equations (5) and (7). It is seen that the results of the experiments agree reasonably well with the latter equation.

(5) CONCLUSIONS

In general, the experimental results discussed in Section (4) seem to lend support to the theories put forward in Section (2). Consequently, it may be assumed (i) that the currents in a transmitting antenna are practically zero at the free ends, with nodes and antinodes spaced alternately along the antenna wire at $\frac{1}{4}$-wavelength intervals, (ii) that the current is sinusoidally distributed along the two portions of a transmitting antenna which are separated by the point of energy input, and (iii) that the currents in the two portions are dependent upon each other in the manner indicated by equations (3) and (4).

In the case of an isolated receiving antenna energized by a uniform radiation field, the current is distributed approximately as indicated by equation (5), or more accurately by equation (6), and current nodes occur at the free ends and at intervals of one whole wavelength measured from both ends of the antenna.

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MULTIPLE REFLECTIONS BETWEEN TWO TUNED RECEIVING ANTENNAE[*]

By Professor L. S. PALMER, D.Sc., Ph.D., Member, W. ABSON, and R. H. BARKER.

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SUMMARY

A consideration of previous work on the reflecting action of an idle or parasitic receiving antenna suggested the desirability of investigating the extent to which multiple to-and-fro reflections between parallel receiving antennae affect the currents produced in them. In order to avoid the possible errors arising from the assumption that tuned half-wave antennae act as Hertzian dipoles, the values of the mutual impedances and phase differences deduced by Carter and tabulated by Tani are used in the present work.

By considering the effects of multiple reflections between two tuned receiving antennae, formulae for the antenna currents are obtained in terms of the currents produced in the antennae when isolated, the distance between the antennae, the wavelength, and the orientation of the plane containing the antennae to the direction of wave propagation.

The validity of these formulae was tested by experiments with electromagnetic waves 80 cm. in length. The experimental results are compared with the graphical representations of the formulae, and for antenna spacings greater than about one-quarter of the wavelength, it is concluded that (i) the effects of multiple reflections do not appreciably affect the optimum or critical spacings between the antennae, but must be taken into account when determining the antenna currents, and (ii) tuned antennae may be considered to act as Hertzian dipoles.

With parallel tuned antennae oriented with the line joining them parallel to the direction of wave propagation, the optimum distances between them for maximum current in the leading antenna are $0{\cdot}31\lambda$, $0{\cdot}85\lambda$, $1{\cdot}36\lambda$, $1{\cdot}87\lambda$, etc. In the trailing antenna the current gradually increases as the antenna spacing increases above a value somewhat less than $0{\cdot}2\lambda$ For maximum current in either antenna when oriented with the line joining them perpendicular to the direction of wave propagation, the optimum spacings are $0{\cdot}67\lambda$, $1{\cdot}71\lambda$, etc., and for other orientations the current variations with antenna spacing are determined by equations (4) and (6) when the field is uniform, and by equations (5) and (7) when the field is not uniform.

(1) INTRODUCTION

The advent of short-wave beam antenna systems has led to the publication of many papers on the radiation impedance and directional properties of various kinds of antenna arrays consisting of two or more parallel wires. Since the early work of Abraham[†] and others, perhaps the most important contributions to the subject are those of Bontsch Bruewitsch,[‡] Wilmotte and McPetrie,[§] Pistolkors,[||] Beckmann,[¶] Carter,[**] Tani,[††] Brown and King,[†] and Brown[‡] all of whom deal essentially with questions of radiation impedance and its dependence on antenna spacings, current distributions and phase differences. In practice, it is usual with transmitting antennae to determine the antenna spacing appropriate to a specified phase difference and to space and feed the antennae accordingly. With a tuned idle or parasitic reflecting antenna (whether used in a transmitting or receiving array) the phase difference is automatically determined by the ratio of the distance between the antennae to the length of the wave. This was discussed by Englund and Crawford[§] in America and by Palmer and Honeyball[||] independently in August 1929. The object of the latter authors was to determine the optimum spacing between two parallel tuned receiving antennae. The theoretical treatment was approximate because, firstly, the antennae were assumed to act as Hertzian dipoles, and, secondly, when the current in either antenna is represented mathematically by an infinite series of terms which may be assumed to represent the effects of an infinite number of to and fro reflections between the antennae, then the previous treatment only considered the first term of the series. In other words, the initial effect of one antenna upon the other was alone considered and not the effect of multiple reflections between them. These multiple reflections, although of progressively decreasing importance, do affect the value of the currents produced in the antennae.

Carter[¶] considered the finite length of each antenna and deduced expressions for the radiation impedance and for the phase difference between the antenna currents. Tani[**] and Brown and King[††] developed similar theories, and the former tabulated the values of the mutual resistance and reactance for various antenna spacings and lengths. Because it seems that Carter's assumptions approximate more closely to the actual experimental conditions than the assumptions underlying the dipole theory, Carter's formulae (using Tani's tables) have been adopted in the present work.

None of the authors cited above has treated the problem of two antennae from the point of view of multiple to-and-fro reflections. In the most recent work[‡‡] Brown states "The current in antenna 1 is so weak that the reaction back into the antenna 0 is negligible." This is not necessarily true, because the current in antenna 1 may be nearly as great as that in antenna 0. Consequently it was felt desirable to extend the previous work of Palmer and Honeyball by considering multiple reflections and by adopting the more exact expressions for the mutual impedance and radiation field which have since been developed by Carter, Tani, and Brown and King. It should then be possible, in the mathematical expression for the antenna current, to determine the importance of those terms representing subsidiary reflections by comparing the experimental values of the optimum antenna spacings with the results previously obtained.

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[*] Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 424.

[†] See Reference (1). [‡] *Ibid.*, (2). [§] *Ibid.*, (3). [||] *Ibid.*, (4). [¶] *Ibid.*, (5). [**] *Ibid.*, (6). [††] *Ibid.*, (7).

[†] See Reference (8). [‡] *Ibid.*, (9). [§] *Ibid.*, (10). [||] *Ibid.*, (11). [¶] *Ibid.*, (6). [**] *Ibid.*, (7). [††] *Ibid.*, (8). [‡‡] *Ibid.*, (9).

The method of analysis employed may be regarded as an analytical expedient. It is not suggested that by multiple reflections the currents in the antennae take a finite time to attain their measured values.

(2) THEORETICAL CONSIDERATION

In order to determine the antenna current produced by any given radiation fields, it is obviously necessary to have a knowledge of the antenna impedance and of the factors upon which it depends. This question has already been the subject of much previous work (see above). In order to simplify the experimental work, a direct determination of the antenna impedance was rendered unnecessary by comparing the current produced in two antenna when both are present with the current produced in either antenna when isolated. This procedure was employed by Brown[*] and involves the usual assumption, namely that, in the neighbourhood of one receiving antenna, the component of the field due to the distant transmitter is not appreciably affected by the proximity of the other receiving antenna. The experimental results of most authorities seem to justify this assumption, at least for distances between the antennae which exceed about $0{\cdot}2\lambda$.

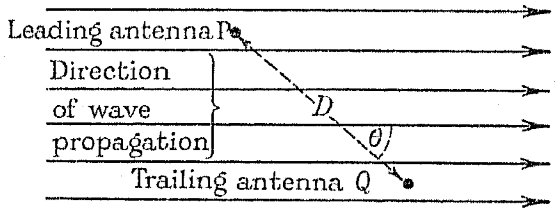

Fig. 1

In what follows the " dashed " symbols (e.g. I') refer to the antennae when isolated and the " undashed " symbols to the antennae when both are present.

In Fig. 1, P and Q are two identical tuned half-wave linear antennae situated in a uniform electromagnetic field polarized with its electric vector parallel to the antennae. This field is produced by some distant transmitter T.

Suppose the antennae are a distance D apart and the plane containing them is oriented at an angle θ to the direction of wave propagation.

Then, with antenna Q (say) absent, let the current produced in the antenna P be ${}_{P}I' = {}_{P}I'_0 \sin \omega t$. At a distance D from P the field due to the re-radiation from P will be proportional to $A\,{}_{P}I'_0 \sin(\omega t - \beta)$, where, for a given antenna, A and β are functions of the distance D and the wavelength λ. According to the dipole theory of Hertz, these constants are given by[†][‡]

$$A = \left(\frac{1}{a^2} - \frac{1}{a^4} + \frac{1}{a^6}\right)^{\frac{1}{2}} \text{ and } \beta = \left(a - \text{arc tan}\,\frac{a^2 - 1}{a}\right)$$

where $a = 2\pi D/\lambda$. Pistolkors's[*] expressions for A and β reduce to $A = (a^2 + \pi^2/4)^{-\frac{1}{2}}$ and $\beta = (a^2 + \pi^2/4)^{+\frac{1}{2}})$ whilst Tani[†] gives tables of values of A and β for different ratios of D/λ in terms of R_{PQ} and X_{PQ}—the radiation resistance and reactance respectively of the aerial array PQ. In Carter's[‡] and Tani's[§] papers

$$A = (R_{PQ}^2 + X_{PQ}^2)^{\frac{1}{2}} \text{ and } \beta = \text{arc tan}\,(-\,X_{PQ}/R_{PQ})$$

If now the antenna Q be placed near P, it will be influenced both by the field from the distant transmitter T and by the re-radiated field from P. That is, the current in Q from these two component fields will be given by the vector sum

$${}_{Q}I'_0 \sin(\omega t - \alpha) + AB\,{}_{P}I'_0 \sin(\omega t - \beta)$$

where ${}_{Q}I'_0$ is the amplitude of the current produced in the antenna Q by the field from the transmitter T, and α is the phase difference $2\pi D \cos\theta/\lambda$. In the second term, B is a constant involving the resistance of the antenna Q, and the wavelength, but is independent of the distance D. In addition to these two component fields in the neighbourhood of Q, there will be other component fields of less magnitude due to the re-radiation from P of energy it has received from Q. The initial current ${}_{Q}I'_0 \sin(\omega t - \alpha)$ in Q will produce a current $AB\,{}_{Q}I'_0 \sin(\omega t - \alpha - \beta)$ in P, which, in turn, will re-radiate to Q, producing a current

$$A^2B^2\,{}_{Q}I'_0 \sin(\omega t - \alpha - 2\beta).$$

Thus the final current ${}_{Q}I$ in Q may be considered to be the resultant of an infinite series of component currents, i.e.

$$\begin{aligned}{}_{Q}I = {}_{Q}I'_0 \sin(\omega t - \alpha) + AB\,{}_{P}I'_0 \sin(\omega t - \beta) \\ + (AB)^2\,{}_{Q}I'_0 \sin(\omega t - \alpha - 2\beta) + (AB)^3\,{}_{P}I'_0 \sin(\omega t - 3\beta) \\ + (AB)^4\,{}_{Q}I'_0 \sin(\omega t - \alpha - 4\beta) + \ldots \quad . \quad . \quad . \quad (1a)\end{aligned}$$

Similarly

$$\begin{aligned}{}_{P}I = {}_{P}I'_0 \sin \omega t + AB\,{}_{Q}I'_0 \sin(\omega t - \alpha - \beta) \\ + (AB)^2\,{}_{P}I'_0 \sin(\omega t - 2\beta) + (AB)^3\,{}_{Q}I'_0 \sin(\omega t - \alpha - 3\beta) \\ + (AB)^4\,{}_{P}I'_0 \sin(\omega t - 4\beta) + \ldots \quad . \quad . \quad . \quad (1b)\end{aligned}$$

These series may be summed quite readily in the special case when $\theta = 90°$, that is, when the plane containing the antennae is perpendicular to the direction of wave propagation, but the exact summations for any other values of θ (or α) are somewhat cumbersome.

When $\theta = 90°$, we have the simplifications:—

$$\alpha = 0, \; {}_{P}I'_0 = {}_{Q}I'_0 = I'_0, \text{ and } {}_{P}I_0 = {}_{Q}I_0 = I_0$$

Hence the amplitude I_0 of the current in either antenna in the presence of the other is given by

$$I_0 = I'_0\left[1 - 2AB\cos\beta + (AB)^2\right]^{-\frac{1}{2}} \quad . \quad . \quad (2)$$

In this equation I_0 and I'_0 can be directly measured, whilst A and β are known constants. B can thus be determined.

In a non-uniform radiation field the second two simplifications may not be justified even when $\theta = 90°$. In this case, let ${}_{Q}I'_0/{}_{P}I'_0 = C$, where C is not now equal to unity. Then equation (2) must be replaced by the equations

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[*] See Reference (9).

[†] See p. 1046 of Reference (11).

[‡] See p. 387 of Reference (12).

[*] See p. 566 of Reference (4).

[†] See pp. 23 and 73 *et seq.* of Reference (7).

[‡] See pp. 1010 and 1016 of Reference (6).

[§] See Reference (7).

$$_PI_0 = {}_PI_0'[1 - 2ABC\cos\beta + (ABC)^2]^{-\frac{1}{2}} \quad . \quad (3a)$$

and $$_QI_0 = {}_QI_0'\left[1 - 2\frac{AB}{C}\cos\beta + \left(\frac{AB}{C}\right)^2\right]^{-\frac{1}{2}} \quad . \quad (3b)$$

Thus, if $_PI_0$ and $_PI_0'$ be measured directly and the value of $({}_PI_0^2 - {}_PI_0'^2)/{}_PI_0^2$ be calculated, the validity of equation (3a) may be tested by the linearity of the graph obtained by plotting the above function of the currents against the expression $[2AC\cos\beta - B(AC)^2]$, treating the latter expression as a function of C and D/λ. The value of the constant B can be determined from equation (2) or from equations (6) and (7) (see below). In a similar way equation (3b) may also be tested.

For values of θ other than 90°, the right-hand sides of equations (1) may be summed approximately by making the assumption that terms involving $(AB)^3$ and higher powers can be neglected. This is equivalent to neglecting all reflections after the first two, a procedure which is justified because experiments have shown that $(AB)^3 < 0{\cdot}01$. Consequently, if the antennae are situated a distance from the transmitter which is very great compared with the distance between them and the radiation field is assumed to be constant over the distance PQ (Fig. 1), then $_PI_0' = {}_QI_0' = I_0'$ and

$$_PI_0 \simeq I_0'[1 + 2AB\cos(\alpha+\beta) + (AB)^2(1 + 2\cos 2\beta)]^{\frac{1}{2}} \quad (4a)$$

and

$$_QI_0 \simeq I_0'[1 + 2AB\cos(\alpha-\beta) + (AB)^2(1 + 2\cos 2\beta)]^{\frac{1}{2}} \quad (4b)$$

If the radiation field in which the antennae are situated is not uniform then $_PI_0' \neq {}_QI_0'$ and equations (4) must be replaced by

$$_PI_0 = {}_PI_0'[1 + 2ABC\cos(\alpha+\beta) + (AB)^2(C^2 + 2\cos 2\beta)]^{\frac{1}{2}} \quad . \quad . \quad (5a)$$

and

$$_QI_0 = {}_QI_0'\left[1 + 2\frac{AB}{C}\cos(\alpha-\beta) + (AB)^2(C^{-2} + 2\cos 2\beta)\right]^{\frac{1}{2}} \quad . \quad . \quad (5b)$$

where $C = {}_QI_0'/{}_PI_0'$ as before. These equations reduce to equations (3) when $\alpha = 0$.

In order to test experimentally the validity of equations (4), it is useful to combine them together, with the consequent elimination of the terms involving $(AB)^2$. We then get

$$\frac{{}_QI_0^2 - {}_PI_0^2}{I_0'^2} = 4AB\sin\alpha\sin\beta \quad . \quad . \quad (6)$$

Similarly equations (5) become

$$\frac{({}_QI_0^2 - {}_PI_0^2) - ({}_QI_0'^2 - {}_PI_0'^2)}{{}_PI_0'\,{}_QI_0'} = 4AB\sin\alpha\sin\beta \quad . \quad (7)$$

In equations (6) and (7) all the currents can be measured and the appropriate function can be plotted against $\sin\alpha[= f(\theta)]$, keeping D/λ and therefore A and $\sin\beta$ constant. Thus the dependence of the currents upon the orientation θ of the plane containing the antennae to the direction of wave propagation may be tested by the linearity of the resulting graph, and the constant B can be determined from its slope.

(3) EXPERIMENTAL WORK

(a) Apparatus

In order to test the foregoing conclusions concerning the values of the currents in two parallel tuned antennae, it was decided to work on ultra-short waves in order to reduce the size of the necessary apparatus to manageable dimensions. For this reason a valve oscillator was used which operated on a wavelength range of from 60 to 100 cm.* The radiating circuit consisted of a half-wave antenna placed at the focus of a cylindrical parabolic reflector of $2\frac{1}{2}\lambda$ aperture and $\frac{3}{4}\lambda$ focal length. The antenna was fed by feeder lines one wavelength long which protruded $\frac{1}{4}\lambda$ through the back of the reflector and the whole apparatus was elevated about $2\frac{1}{2}$ m. (or about 3λ) above the ground, in order to minimize ground reflection effects. From previous experiments† in the locality, reflections from the ground were found to be negligible at heights greater than about $2\frac{1}{2}\lambda$.

The receiving antennae consisted of short brass rods which were mounted vertically on a wooden arm capable of rotating in a horizontal plane about a vertical axle through its centre. The revolving arm was at such a height above the ground that the antennae were level with the transmitter. The two antennae could slide along the arm, thereby changing the distance D between them, whilst the rotation of the arm controlled the value of θ. The currents were measured by inserting vacuum thermojunctions at the centre of the antennae, and arrangements were made for the d.c. leads to remain perpendicular to the antennae as they rotated. In order to get a tuned half-wave antenna, the length of the rod was gradually adjusted until the current recorded was a maximum. The geometrical length was then found to be about 19 cm. shorter than $\lambda/2$. This difference in length agreed with the similar value determined by Palmer and Gillard‡ using the same wavelengths with similar apparatus.

(b) Experimental Results

When an isolated antenna was moved about in the locality in which the experiments with two antennae were to be carried out, it was found that the current in the isolated antenna did not remain constant. No adjustment of the reflector or transmitting aerial would make the field sufficiently uniform. Owing to this inability to obtain a sufficiently uniform field throughout the region in which the receiving antennae were situated, it was not possible to check directly the validity of equations (2), (4), and (6). These equations assume $_PI_0' = {}_QI_0'$ (or $C = 1$). Consequently values of $_PI_0'$ and $_QI_0'$ were recorded independently and experiments were undertaken to test the validity of equations (3), (5), and (7), which do not assume that $_PI_0' = {}_QI_0'$.

(i) Antenna oriented so that θ = 90°.

The receiving antennae were set up at a distance of about 6 or 8 m. from the transmitter, so that the plane

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* The details of this oscillator are described by Palmer and Gillard in their paper on page 285.

† See page 70 of Reference (14). ‡ See page 421 of Reference (13).

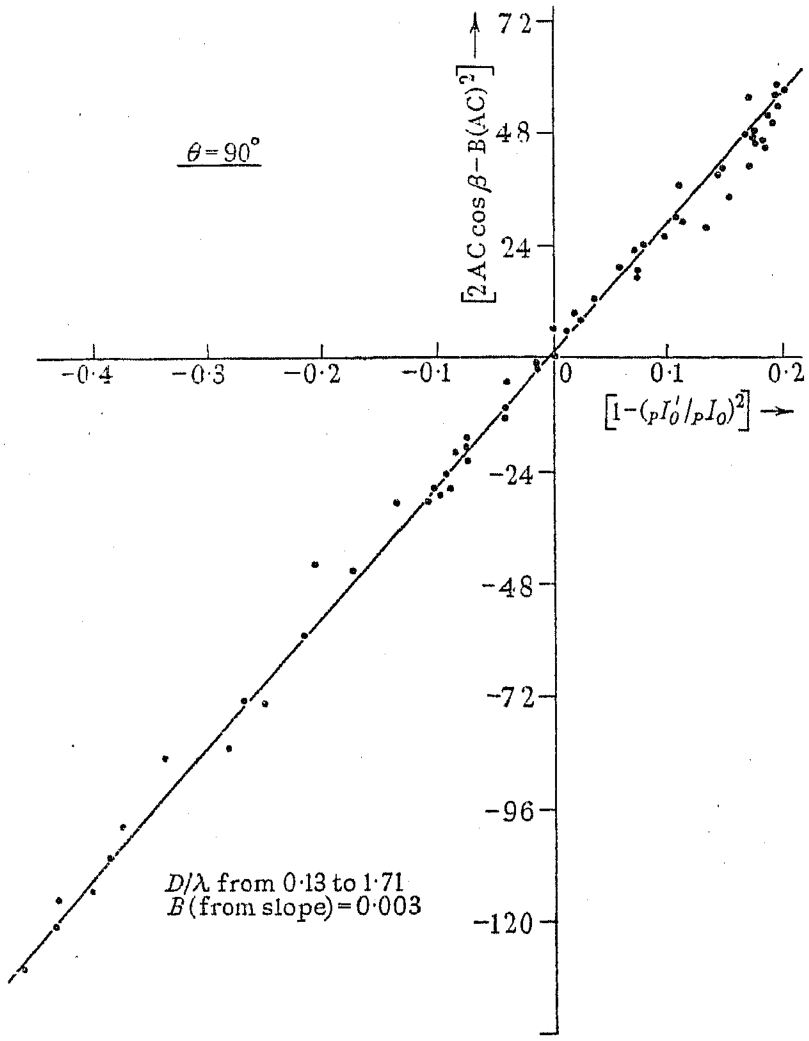

Fig. 2

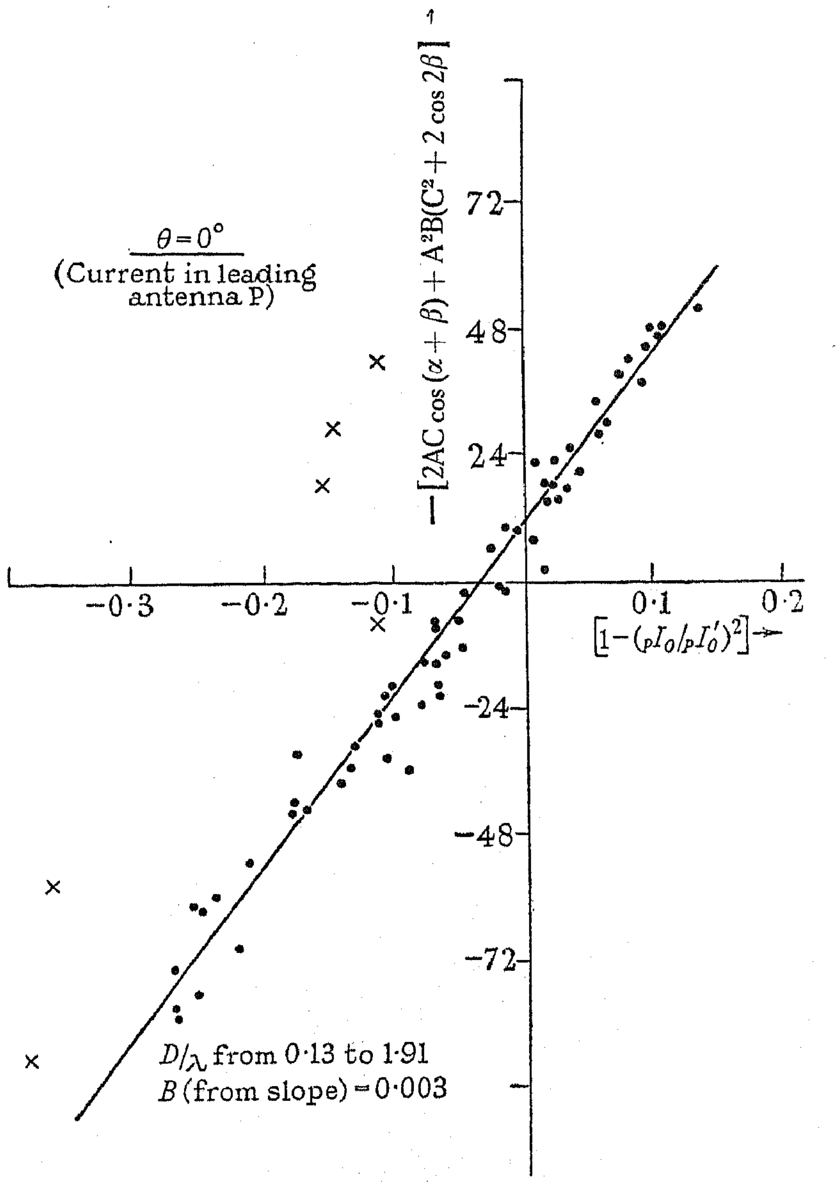

Fig. 4

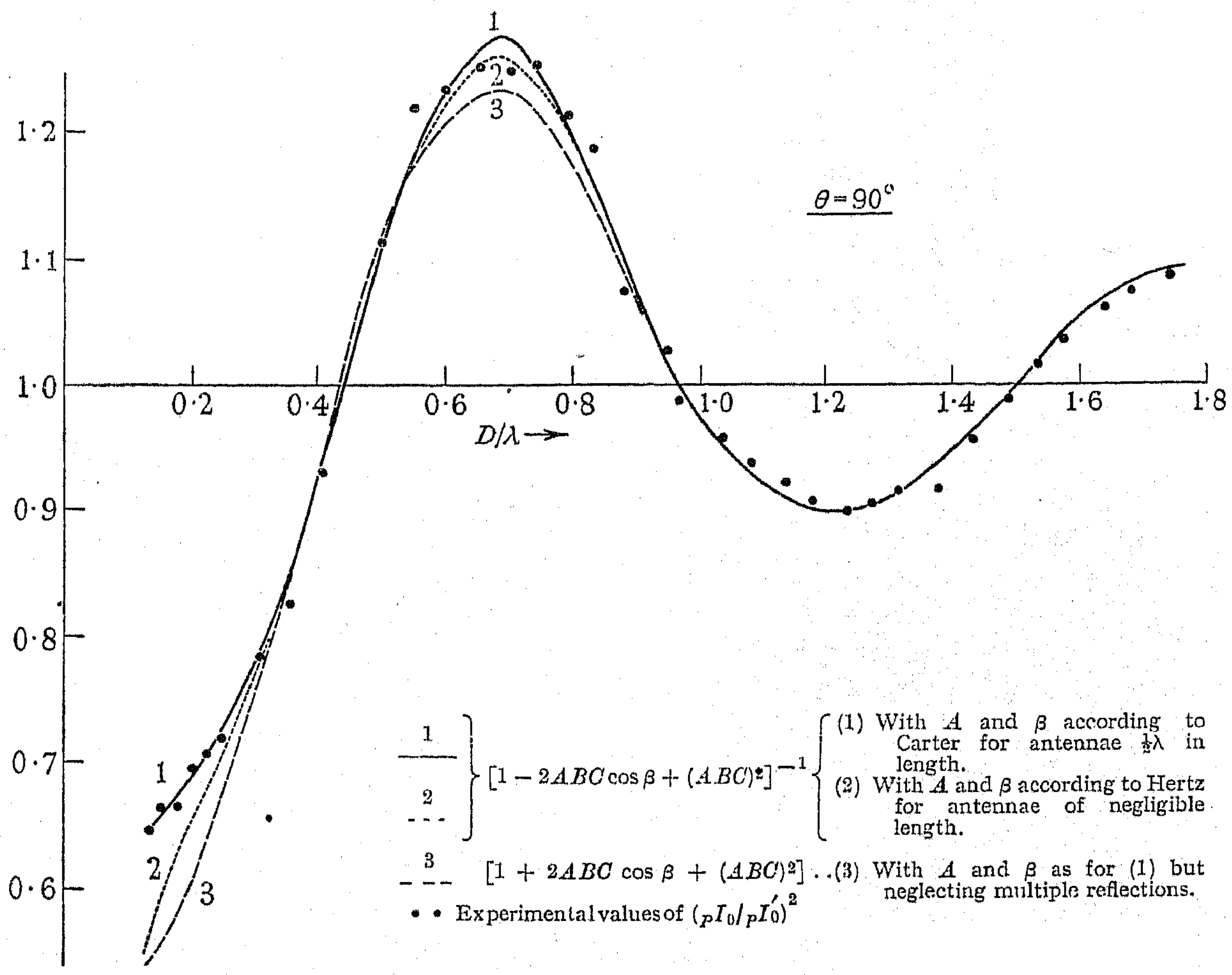

Fig. 3

containing the antennae was perpendicular to the direction of wave propagation ($\theta = 90°$). Measurements of ${}_PI_0$ (or ${}_QI_0$) were taken for 62 values of D/λ, ranging from 0·13 to 1·71, and after each observation one antenna was removed and the value of ${}_PI_0'$ (or ${}_QI_0'$) was then read. The function $[1-({}_PI_0'/{}_PI_0)^2]$ plotted against $[2AC\cos\beta - B(AC)^2]$ is shown in Fig. 2, which is a linear graph to within the errors of experiment. In this graph there is no indication of the values of D/λ for which the current ${}_PI_0$ (or the ratio ${}_PI_0/{}_PI_0'$) is a maximum or a minimum. Consequently the same data have been replotted in Fig. 3 so as to show the variation of the current ${}_PI_0$ with increasing values of the ratio D/λ. The full-line graph is that of equation (3*a*) in which the values of A and β in the expression $[1-2ABC\cos\beta + (ABC)^2]^{-1}$ have been calculated from Carter's theory, using Tani's tables for antennae of length equal to $\lambda/2$. The dotted graph is also from equation (3*a*), but the values of A and β are those used by Palmer and Honeyball, based on Hertz's dipole theory which assumes the antennae to be of negligible length (see page 295). The " dashed " graph neglects multiple reflections, and the ordinates have been calculated from $[1 + 2ABC\cos\beta + (ABC)^2]$, which expression is obtained by taking only the first two terms of the original series.

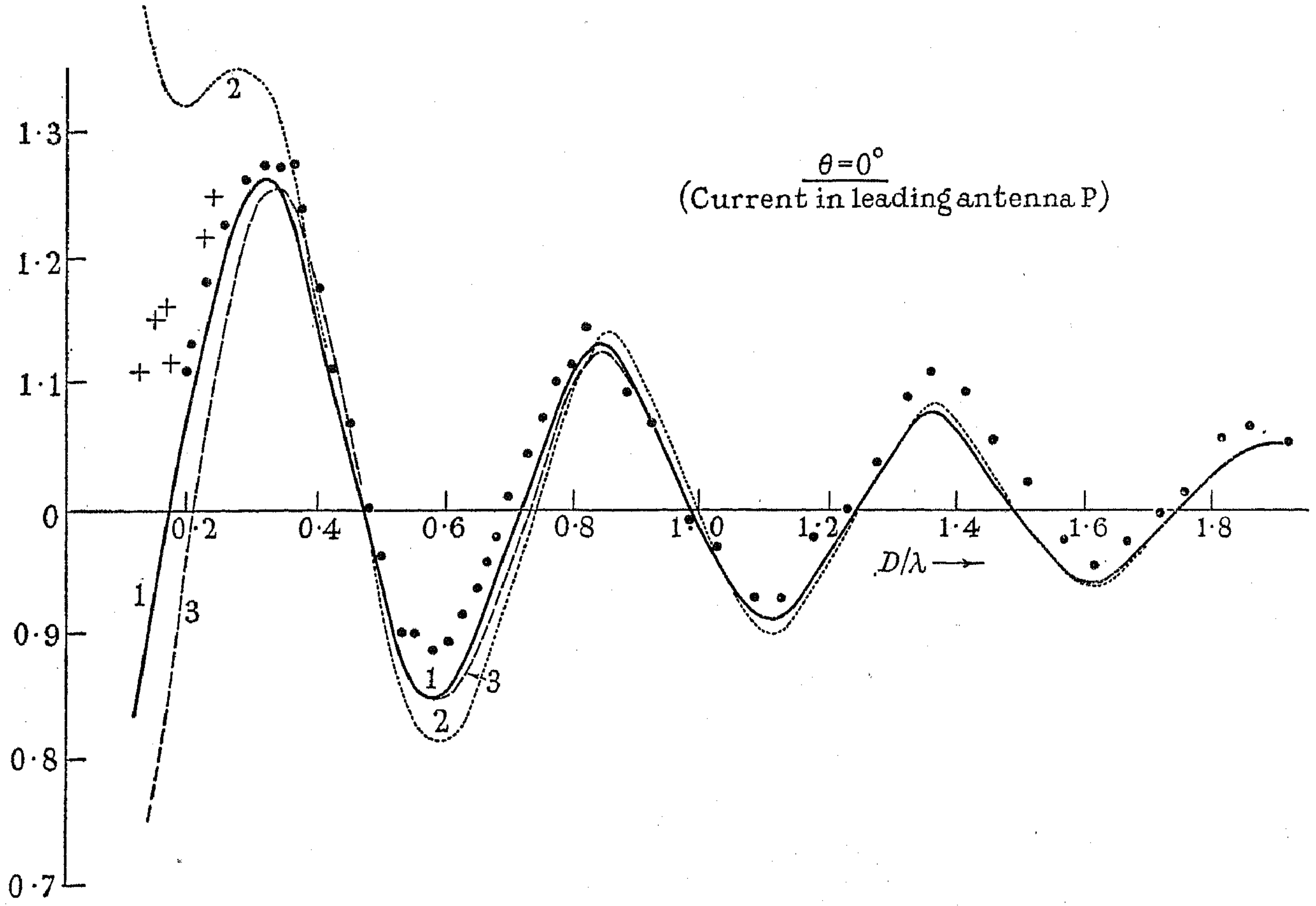

Fig. 5

1, 2 } $[1 + 2ABC\cos(\alpha+\beta) + (AB)^2(C^2 + 2\cos 2\beta)]$ { (1) With A and β according to Carter for antennae $\frac{1}{2}\lambda$ in length. (2) With A and β according to Hertz for antennae of negligible length.

3 $[1 + 2ABC\cos(\alpha+\beta)]$ . . . . (3) With A and β as for (1) but neglecting multiple reflections.

• • + + }Experimental values of $({}_PI_0/{}_PI_0')^2$

In Fig. 3 it may be noted that the current ratios for values of D/λ as small as 0·13 agree to within the errors of experiment with the theoretical full-line graph. This is in contrast with the lack of agreement for such small values of D/λ when the plane containing the antennae is not perpendicular to the direction of wave propagation and $\theta \neq 90°$ (see Figs. 5, 7, and 9).

(ii) Antennae oriented so that θ = 0°.

Similar experiments were carried out with the receiving antennae and transmitting aerial all in line. Sixty-seven measurements of ${}_PI_0/{}_PI_0'$ for values of D/λ ranging from 0·13 to 1·91, and 33 measurements of ${}_QI_0/{}_QI_0'$ for values of D/λ from 0·13 to 1·27, were taken. The data are plotted as linear graphs in Figs. 4 and 6 and with D/λ as abscissae in Figs. 5 and 7 respectively.* In these latter figures the full-line curves are the graphical representations of equations (5), in which Tani's values of A and β are used; and the dotted graphs are also calculated from equations (5) but with the values of A and β deduced from the dipole theory of Hertz. The " dashed " graphs are based on the assumption that multiple reflections are negligible in their effect on the currents ${}_PI$ or ${}_QI$.

Because of the curious deviation of certain points (marked with a cross in Figs. 4, 5, 6, and 7) from the theoretical curves for values of D/λ less than about 0·2, it was thought advisable to check the data for this orientation of the antennae and plot the results in accordance with equation (7) in which, since $\theta = 0°$, the right-hand side reduces to $4AB \sin(2\pi D/\lambda) \sin\beta$. It is significant that no such deviation occurs for small values of D/λ in Figs. 2 and 3, for which $\theta = 90°$. The 12 points marked with a cross in Fig. 8 were obtained for values of D/λ ranging from 0·13 to 0·30. The other 21 points vary from D/λ equal to 0·30–0·80. The curious deviations noted above for small values of D/λ are here definitely confirmed.

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* Owing to a calibration variation the value of ${}_QI_0'$ was found to be 2 % too high. The experimental points in Figs. 6 and 7 have been corrected accordingly.

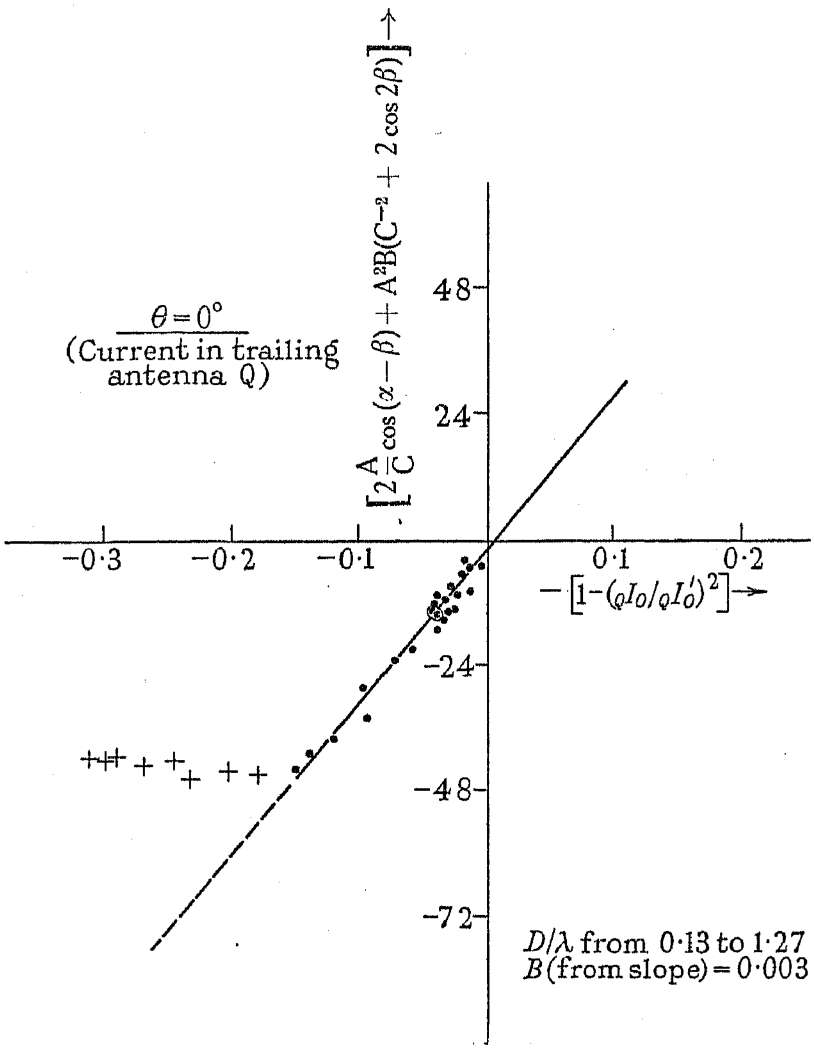

Fig. 6

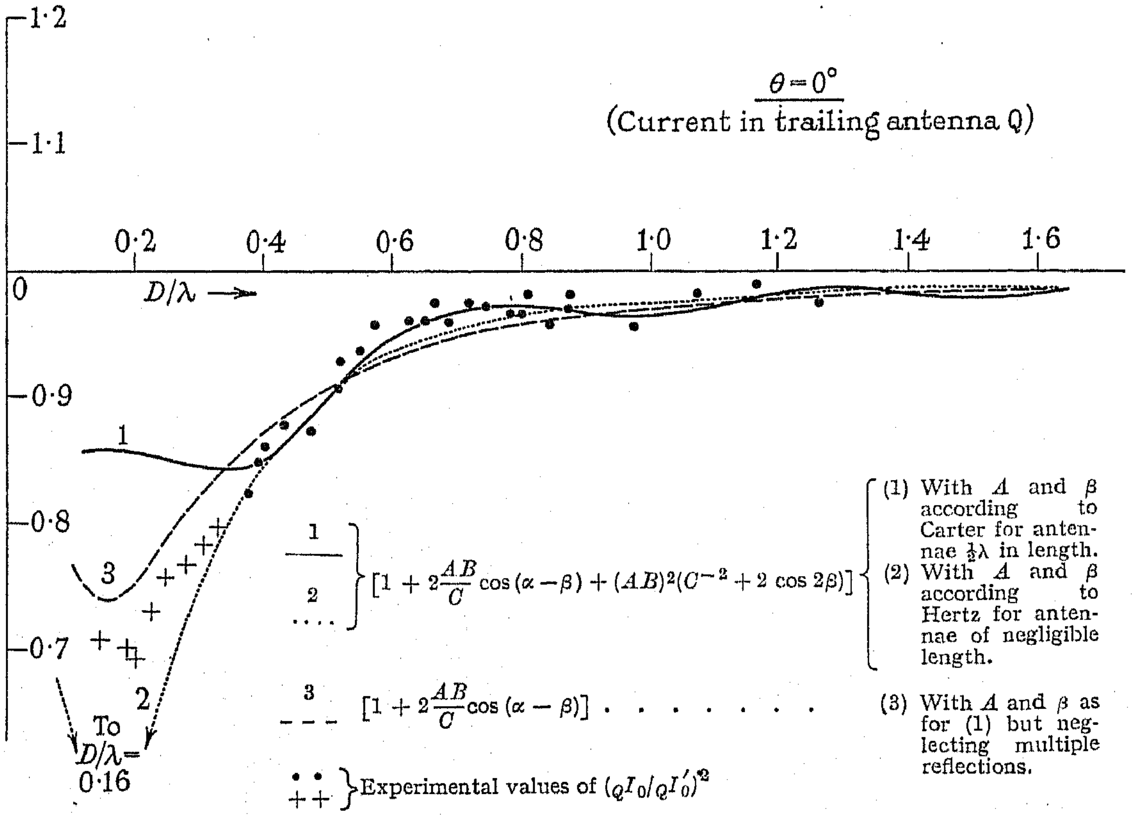

Fig. 7

(iii) Antennae with the plane containing them oriented at any angle to the direction of wave propagation.

Finally, experiments with the antennae revolving (variable θ) were carried out for four constant values of D/λ, viz. 0·50, 0·42, 0·38, and 0·22. The results are calculated according to equation (7) (using $B = 0{\cdot}003$ as deduced from Figs. 4 and 6) and are shown as a linear graph in Fig. 9 and as polar diagrams in Fig. 10. In this latter figure the radii vectores are proportional to the current amplitude ${}_PI_0$. The signficance of the divergence from the theoretical graphs of most of the experimental points marked with a cross, for which D/λ is equal to 0·22, is discussed below.

(4) CONCLUSION

From the linearity of the graphs shown in Figs. 2, 4, 6, 8, and 9, it may be concluded that the currents in two parallel tuned receiving antennae would vary with the antenna spacing (i.e. with D/λ) and with the orientation of the plane containing the aerials (i.e. with θ) in the manner indicated by equations (2), (4), and (6), when situated in a uniform radiation field, and vary in accordance with equations (3), (5), and (7), when in a non-uniform field. But from the deviations for small values of D/λ shown by the points marked with a cross in Figs. 4 to 8 for which $\theta = 0°$, it is evident that the theoretical equations are not applicable to values of D/λ less than about 0·25 when the line joining the antennae passes through the transmitter. Furthermore, by considering the divergence of the points marked with a cross in Figs. 9 and 10 it is apparent that the foregoing theory applies to the currents in parallel antennae *which are close together* (i.e. small D/λ) only when the line joining them is more or

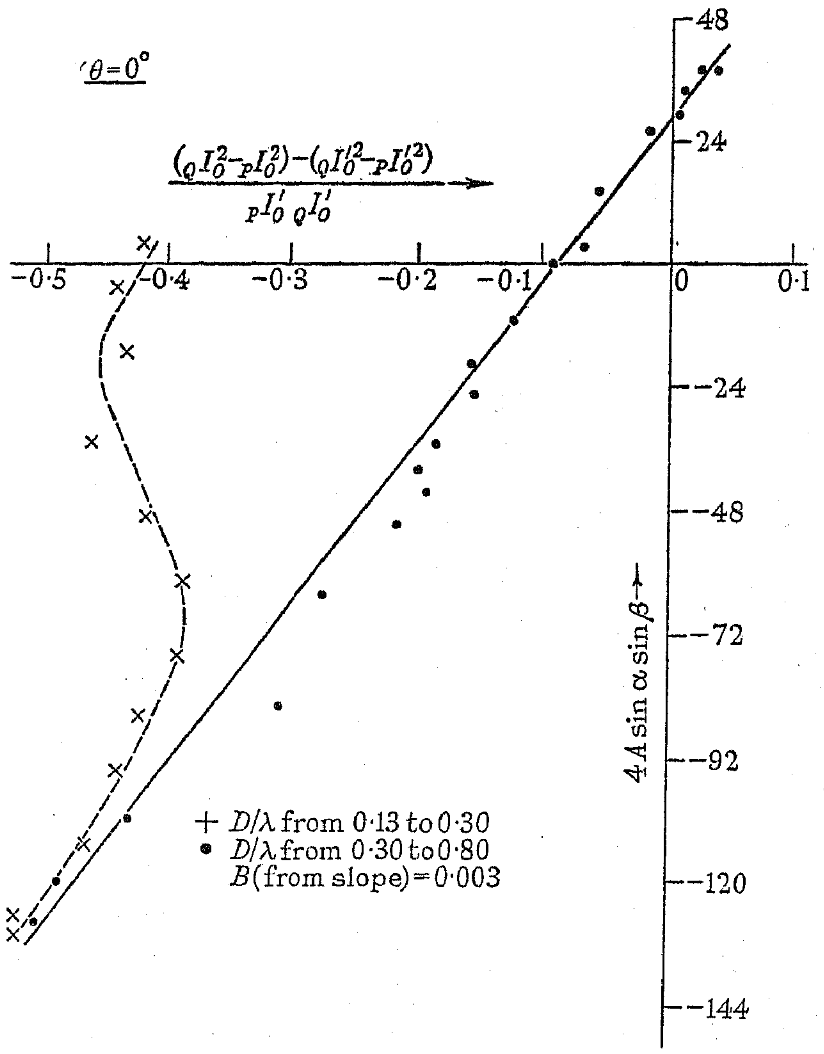

Fig. 8

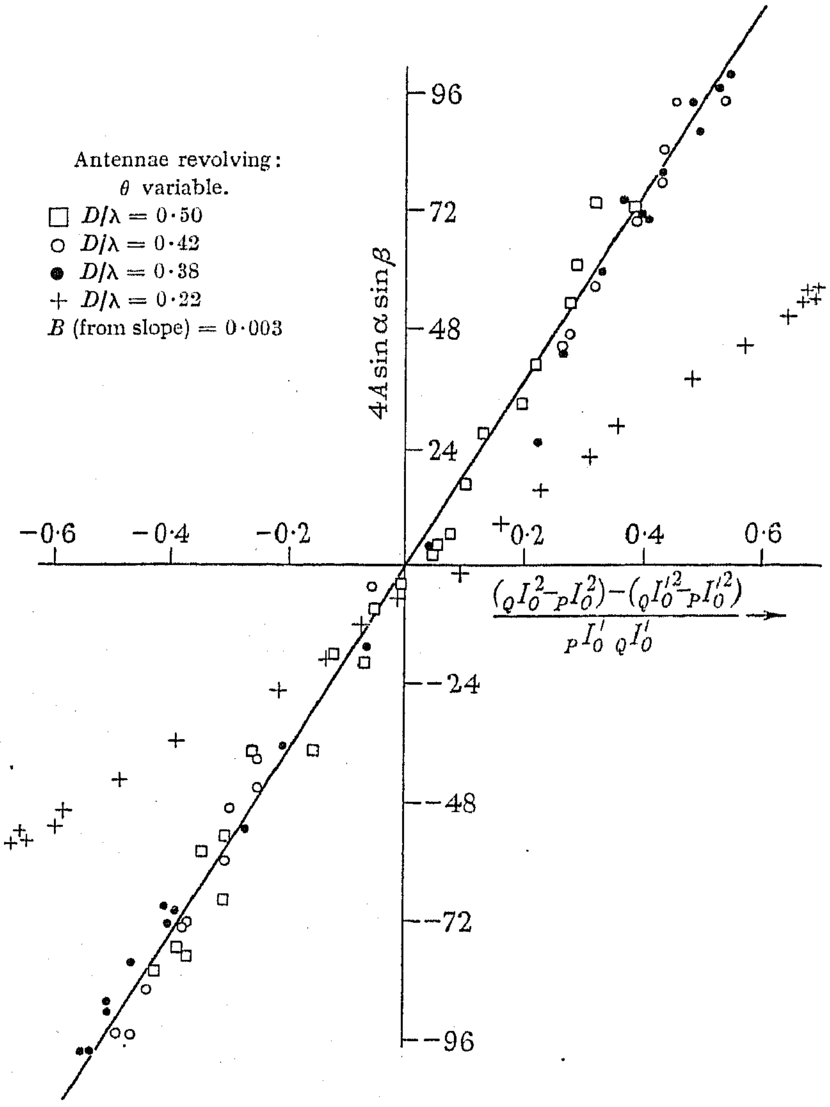

Fig. 9

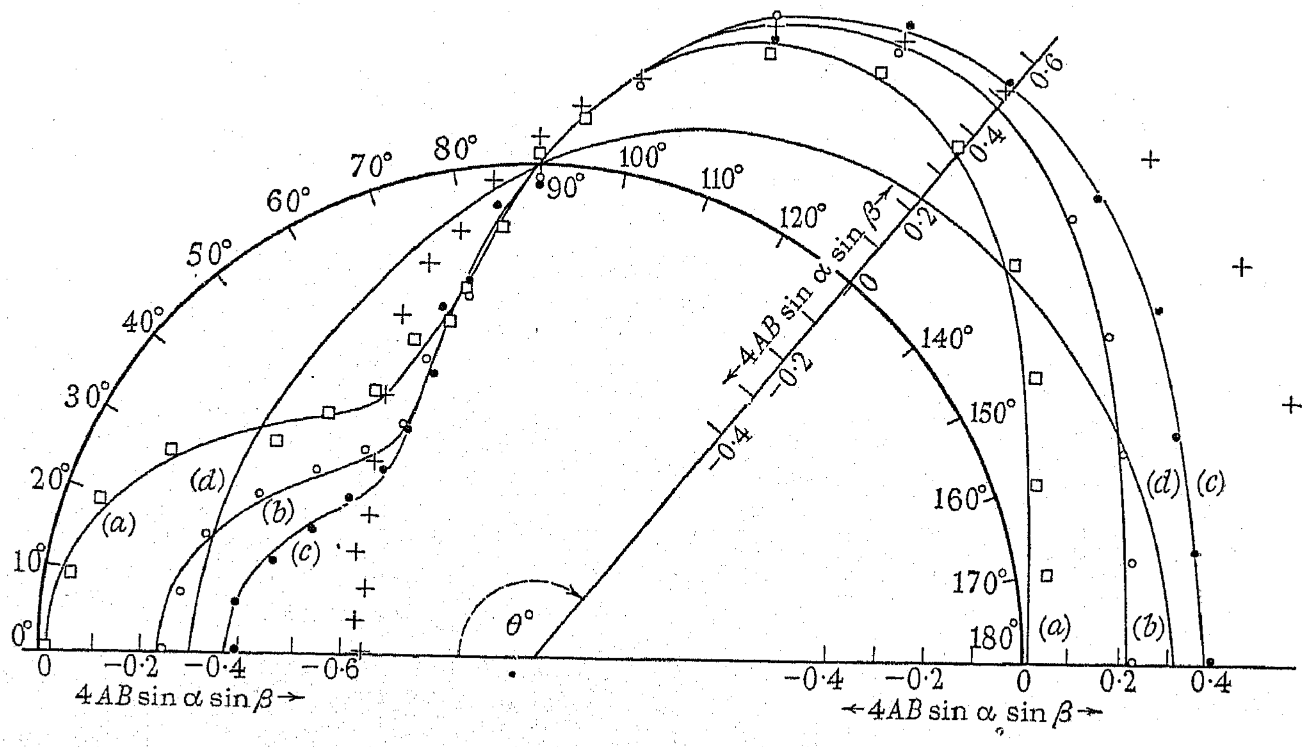

Fig. 10

(a) —— □ —— D/λ = 0·50
(b) —— ○ —— D/λ = 0·42
(c) —— ● —— D/λ = 0·38
(d) —— + —— D/λ = 0·22

Experimental points are:—
$\dfrac{({}_QI_0^2-{}_PI_0^2)-({}_QI_0'^2-{}_PI_0'^2)}{{}_PI_0'\,{}_QI_0'}$

less perpendicular to the direction of wave propagation (i.e. $\theta \simeq 90°$). This follows from the fact that the points marked with a cross in Fig. 9 only coincide with the straight line at the origin, where $\theta = 90°$, whilst in Fig. 10 such points are near the curve (*d*) only in the neighbourhood of 90°. In other words some unconsidered effect is operative when the antennae are close together and the line joining them is not perpendicular to the direction of wave propagation. It seems probable that, with these particular conditions, the assumption that the presence of one antenna does not affect the field produced by the transmitter in the neighbourhood of the other is not true. A comparison of the points marked with a cross in Figs. 5 and 7 shows that the current in the leading antenna P is abnormally increased at the expense of the current in the trailing antenna Q.

Table

CRITICAL VALUES OF D/λ FOR TWO PARALLEL TUNED ANTENNAE

References:	Englund and Crawford* 1929	Palmer and Honeyball† 1929	Carter‡ 1932	Palmer and Taylor§ 1935	Brown‖ 1937	Figs. 5 and 3 of present paper 1937	
For max. current	0·32 0·85	0·33 0·86	— —	0·316 0·856	0·15 0·80	0·31 0·85 1·36 1·87	for $\theta = 0°$
For min. current	0·60	0·60	—	—	0·59	0·58 1·10 1·61	
For max. currents	0·71	0·71	0·69	0·694	0·68 1·75 2·79	0·67 1·71	for $\theta = 90°$
For min. currents	1·22	—	1·31	—	1·21 2·23	1·20	

* See Reference (10). † *Ibid.*, (11). ‡ *Ibid.*, (6). § *Ibid.*, (12). ‖ *Ibid.*, (9).

Fig. 3 shows the optimum antenna spacings for maximum and minimum currents with idle or parasitic antennae, when the plane containing the antenna is perpendicular to the direction of wave propagation; whilst Fig. 5 gives similar information for the leading antenna when the plane containing the antennae is parallel to the direction of wave propagation. The critical values of D/λ from these figures are compared in the Table above with the values obtained by previous authorities.

With the exception of the smallest value of D/λ for $\theta = 0°$, it is apparent that the differences between the various theoretical treatments are, on the whole, less than the errors of experiment when measurements of the critical values of D/λ are made at ultra-high frequencies. Consequently, at these frequencies and with antennae not closer than $0{\cdot}25\lambda$, unless $\theta = 90°$, it is concluded that (i) tuned half-wave receiving antennae may be considered to act as Hertzian dipoles, (ii) multiple reflections have little effect in determining optimum antenna spacings but have a more marked effect on current magnitudes, (iii) either antenna does not appreciably affect the incident field in the neighbourhood of the other, and (iv) the variations in the received currents with changes in the ratio D/λ and with the orientation θ are given by equations (4) and (6) when the radiation field is uniform, and by equations (5) and (7) when the field is not uniform.

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RADIO DIRECTION-FINDING ON WAVELENGTHS BETWEEN 6 AND 10 METRES (FREQUENCIES 50 TO 30 MC./SEC.)*

By R. L. SMITH-ROSE, D.Sc., Ph.D., Member, and H. G. HOPKINS, B.Sc., Ph.D.

[From the National Physical Laboratory.]

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SUMMARY

The following paper describes the development of simple experimental direction-finders for wavelengths between 6 and 10 metres, and their use in an investigation of the accuracy of direction-finding on these wavelengths. It is shown that the inherent accuracy of the instruments is well within the limits of ± 2°, while their sensitivity is sufficient for observation on an experimental 50-watt transmitter at ranges up to 22 miles over flat ground. Bearings have also been taken on signals emanating from London television transmitters, blind-landing beacons, and also from commercial and amateur transmitting stations some 3 000 miles away.

Detailed experiments have shown that the site on which the direction-finder is used must be clear of obstacles, particularly trees and vertical wires, for a radius of at least 50 to 100 yards; and there are indications that similar conditions are necessary for the site of the transmitter. When such conditions are satisfied, the bearings observed at distances up to 22 miles from the transmitter may be in error by as much as 8°, although in the majority of cases the error was less than 2°. Such errors tend to diminish in magnitude as the range increases, and they may or may not be affected by small changes in position of the direction-finder, by changes in frequency of the transmitter, or in the orientation of the transmitting aerial. For a given set of conditions the changes in bearings observed from day to day do not exceed about 2° for ranges of 20 miles, although in some long-distance observations made on signals from American stations at a range of 3 000 miles the variations in bearings were much larger.

A brief study has been made of the behaviour of the loop direction-finder when horizontally polarized waves are emitted at the transmitting station. It is shown that the errors experienced in this case can be almost entirely eliminated by the use of a rotating spaced vertical aerial arrangement in place of the loop. The evidence resulting from the use of this Adcock type of direction-finder indicates that the errors with the loop set are due to the reception of horizontally polarized waves.

(1) INTRODUCTION

Radio direction-finding on ultra-short waves may really be said to have been initiated by Hertz, for in some of his classical experiments carried out some 50 years ago he used circular or rectangular loops of wire with small spark-gaps for detecting the directions of the fields in the waves radiated from his spark-coil-driven oscillators. The wavelength used in many of these experiments was approximately 10 metres or less. From that time very little work appears to have been carried out on the use of direction-finding receivers on wavelengths below 10 metres, although from time to time various systems for transmitting directional and rotatable beams have been employed. In recent years these have led to the development on a commercial scale of the course-indicating beacons and landing-beam systems now in use for the guidance of aircraft.

On wavelengths of about 1 metre S. Uda† has studied the transmitting and receiving properties of half-wave antennae and reflector systems. In addition he has described‡ a direction-finder for a wavelength of 0·50 metre, comprising a receiving antenna at the focal line of a cylindrical parabolic reflector. As this apparatus is rotated during the reception of signal the intensity passes through a minimum at about 35° on either side of its maximum value, and by observation of these minima the direction of the maximum can be located with fair accuracy. Various amateur experimenters have also used the ordinary loop direction-finder on wavelengths of about 5 metres.

The investigation of the performance of the single rotating-loop direction-finder was commenced at the National Physical Laboratory in 1930, and the results of some experiments made on a wavelength of 7·9 metres have been published previously.§ After being in abeyance for some years the study of this subject has recently been resumed, and the present paper describes the results of investigations carried out on wavelengths between 6 and 10 metres during the past two years.

(2) DESCRIPTION OF THE DIRECTION-FINDERS EMPLOYED

The work described in this paper was largely carried out in the field, and the direction-finders were accordingly designed to be of a readily transportable, battery-operated type, with aural indication by means of head-telephones. While instruments of both the rotating-loop and the rotating spaced aerial or Adcock aerial pattern have been employed, attention was first devoted to the loop type on account of its compactness and simplicity.

(a) Rotating-Loop Direction-Finder

The rotating-loop direction-finder depends for its operation upon the fact that the electromotive force induced in a vertical loop by vertically polarized electric waves is proportional to the cosine of the angle between the plane of the loop and the plane of propagation of the waves. Thus when these planes are perpendicular to one another the signal e.m.f. is of zero value, and the practical operation of the direction-finder consists in rotating the loop and determining the position at which the observed signal, detected aurally or visually, passes through a minimum or zero value. The satisfactory operation of a direction-finder in this manner depends upon the attainment of exact electrical symmetry of the loop and associated circuits, and this problem requires great care and attention in the design and construction of apparatus for use at very short wavelengths. In the two classes of loop direction-finder described in this section, this electrical symmetry has been attained by the use of a screened loop connected to a push-pull detector or amplifier circuit.

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* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 87.

† See Reference (1).

‡ *Ibid.*, (2).

§ *Ibid.*, (3).

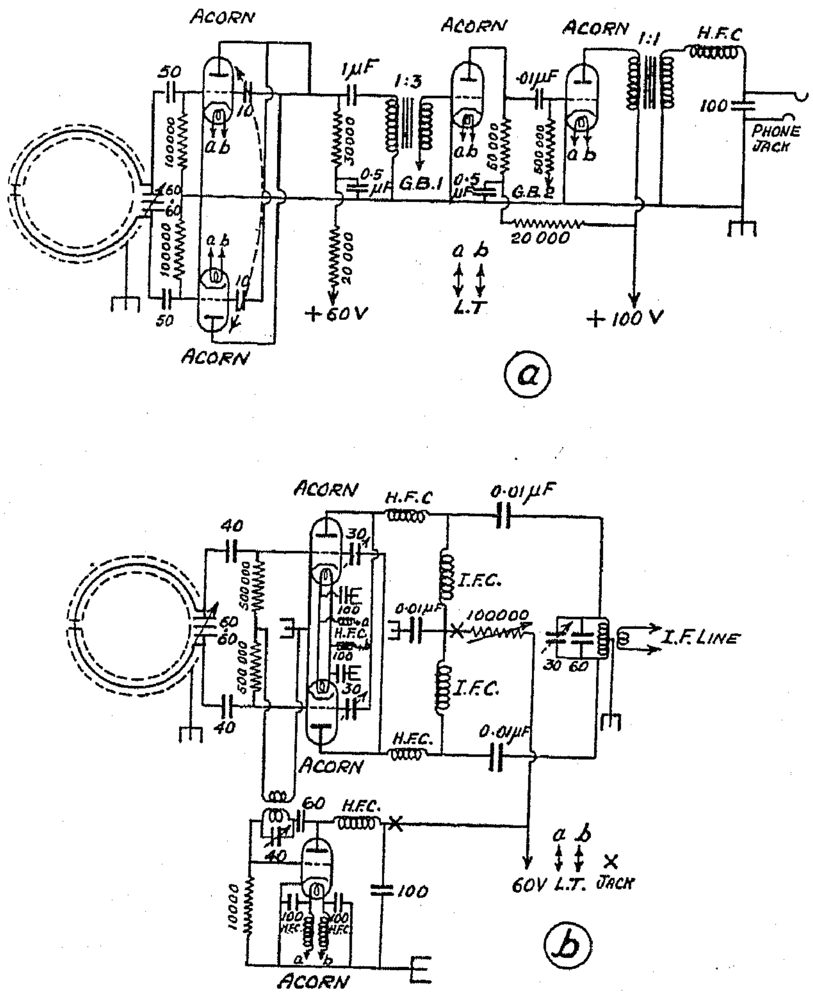

Fig. 1.—Loop direction-finders.

(*a*) Straight receiver.
(*b*) Frequency-changer of superheterodyne.

The capacitance and resistance values are in micro-microfarads and ohms respectively, unless otherwise stated.

The first instrument utilizes a simple straight receiver and has been developed for use in relatively strong fields of the order of 0·1 mV per metre. The circuit diagram is given in Fig. 1(*a*), from which it is seen that the tuned receiving loop is connected to a push-pull detector, with ganged retroaction condensers and a two-stage audio-frequency amplifier. Either modulated or continuous-wave signals can be received. The whole receiver with its batteries is enclosed in a metal box which, together with the screened loop, rotates as a unit on a suitable turntable.

The more sensitive instrument, for use in weaker fields, employs a receiver of the supersonic heterodyne type. In this case the receiver is constructed in two units. The first or rotating unit comprises the frequency-changer in a screened box carrying the screened loop, while the second is a fixed unit containing the intermediate and audio-frequency amplifiers. The two units are coupled together by a screened flexible transmission line carrying the intermediate-frequency output from the frequency-changer.

The circuit diagram of the latter is shown in Fig. 1(*b*). The push-pull detector stage is screened from the local signal frequency oscillator, the output from which is applied in parallel, through a coupling coil, to each of the detectors. Retroaction is applied by means of small pre-set condensers, and is controlled by an adjustment of the anode voltage by means of a variable resistance. The intermediate-frequency output from the detectors is fed to the transmission line through a step-down transformer, which is isolated from the detector anodes by chokes designed to prevent the passage of signal-frequency current.

As already remarked, this transmission line connects the rotating unit of the receiver to the fixed unit containing the intermediate-frequency amplifier, second detector, and audio-frequency amplifier, giving output to the head-telephones. This unit is of a standard pattern and needs little description. An intermediate frequency of about 2 Mc./sec. is used, in order to permit of reasonably high amplification with a large band-width. The latter feature results in an improved performance in the reception of frequency-modulated transmissions, facilitates tuning, and minimizes the need for repeated adjustment due to drift of the frequency of the local oscillator. Provision for the reception of continuous waves is made by the inclusion of a beat-frequency oscillator, which can be switched on as required.

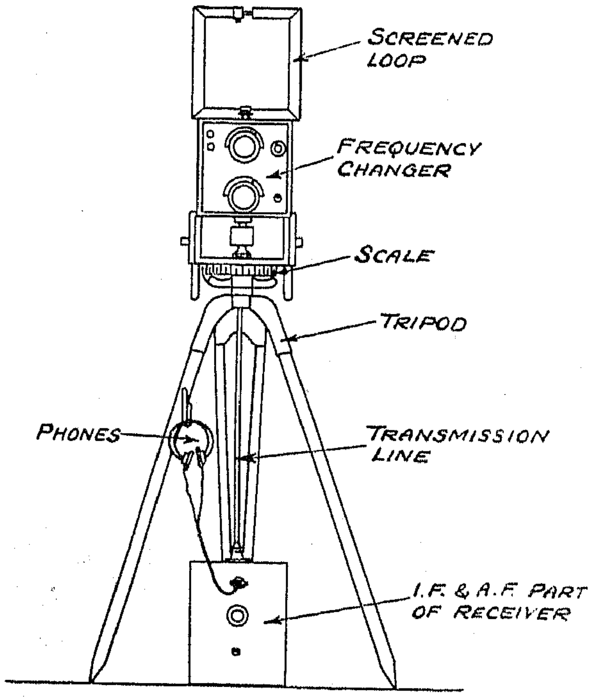

Fig. 2.—Loop direction-finder (super-heterodyne model).

Both the straight and supersonic heterodyne receivers cover a wavelength range of 5·5 to 10·7 metres (frequency 54 to 28 Mc./sec.) by means of two plug-in screened loops. These are square in shape and have sides of $11\frac{1}{2}$ in. and $7\frac{1}{8}$ in. respectively: they are made of $\frac{3}{16}$ in. diameter copper rod surrounded by a 1 in. diameter copper screen. This screen is split at the centre of the top side, and an adjustable collar sliding on the screen provides a convenient means of accurately balancing the input system.

The disposition of the more sensitive direction-finder is shown diagrammatically in Fig. 2. The rotating-loop unit is mounted on a tripod well above the observer, who sits on a stool as nearly as is practicable beneath the centre of the system. With this arrangement the actual position of the observer about the central axis of the direction-finder is not of great importance. The simpler direction-finder is carried in a similar way on a turntable mounted on a tripod, the telephone leads passing through the bearing shaft in place of the transmission line shown in Fig. 2. With each instrument a suitable hand-wheel is provided for rotating the turntable, and a graduated scale is set at a convenient level for the observer to read off bearings from the position of the pointer attached to the rotating portion. In setting up the instruments in the field a magnetic compass is employed to ascertain the direction of the meridian, and suitable sights are provided on the screened loops to facilitate this operation.

The sensitivity of the direction-finder with the straight receiver is such that bearing observations can be made with a swing through the minimum signal position of 5° when operated in a uniform vertically-polarized field, the intensity of which is of the order of 100 microvolts per metre. The more sensitive instrument enables the same observational accuracy to be attained in field intensities of about one-tenth this value. These figures apply to the reception of fully modulated signals. The effective sensitivity in continuous-wave reception is somewhat less than the above, owing to fluctuations of the audio-frequency note due to variations in the local oscillator frequency as the set is rotated. It is more difficult to obtain an accurate reading on this fluctuating note than on a steady modulation note.

(b) Experimental Rotating Spaced Aerial Direction-Finder

It is now well known that one of the limitations of the closed-loop direction-finder is that it is subject to errors in observed bearing when employed in the presence of downcoming horizontally polarized waves. One of the most effective means of overcoming this difficulty is to replace the loop by a pair of spaced vertical aerials in accordance with the arrangement first described by F. Adcock. In Section (5) of this paper some experiments are described in which a source of horizontally polarized waves was employed, and for the purpose of investigating the phenomena observed a simple rotating Adcock direction-finder was developed. It is convenient to outline the main features of this instrument at the present stage.

The construction of this direction-finder involved the replacement of the loop attached to the supersonic heterodyne receiver by a suitable spaced-aerial system with the necessary transmission lines and transformer. The general layout of the instrument is shown in Fig. 3, from which it will be seen that the aerial pair is of the "H" type supported centrally at the vertical axis of rotation. The aerials are made of copper rod, $\frac{1}{4}$ in. diameter and of a total length of 4 ft., and are rigidly attached to wooden supports. Each aerial is supported at its centre by an aluminium tube, $\frac{3}{4}$ in. diameter, forming the screen for the transmission line which consists of two copper rods $\frac{1}{8}$ in. diameter and 3 ft. long. The inner end of each aluminium tube is rigidly attached to the top of the metal frequency-changer box, and the transmission lines are connected inside the box to the primary winding of an air-core screened transformer. The secondary of this transformer is outside the primary and coaxial with it, a screen of copper foil strips being placed between the windings. This secondary winding is tuned and connected to the push-pull detector in the same manner as was the closed loop in the sets previously described.

The satisfactory operation of this spaced-aerial direction-finder depends upon the attainment of symmetry of the two aerials and also of the two halves of each aerial. For this reason care must be taken to set up the direction-finder on a level surface, and also to place the observer close in to the central axis and on the line at right angles to the plane containing the two aerials. The latter precaution necessitates the observer changing his position for each working azimuth; and although in a fixed installation this could probably be avoided by elevating the aerial system, this would appear to be impracticable in a set designed for field use.

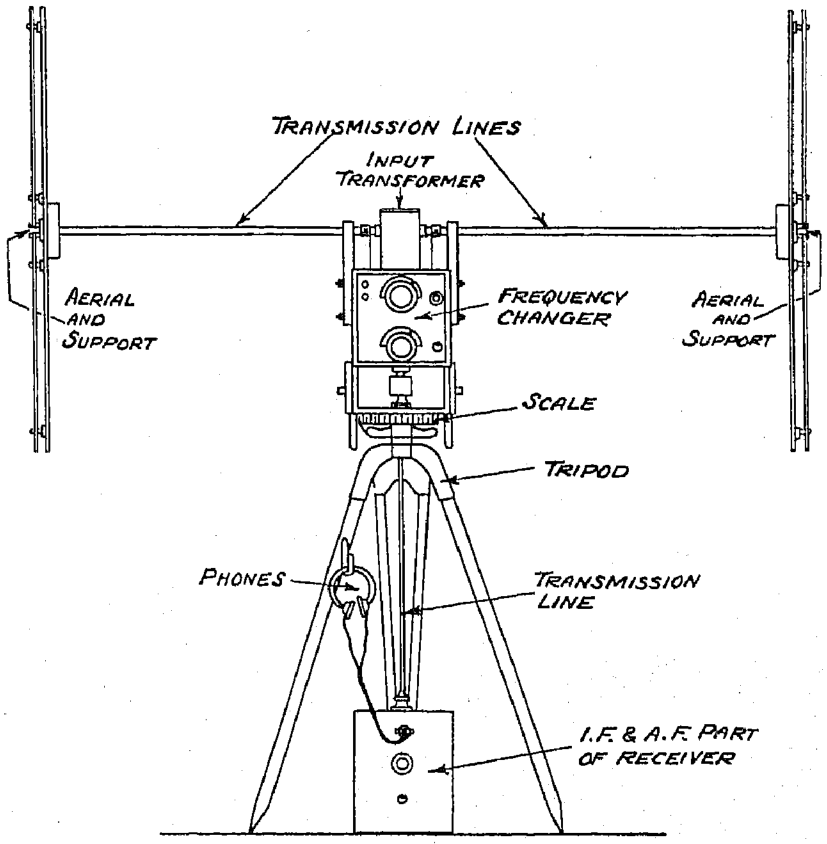

Fig. 3.—Spaced-aerial (Adcock) direction-finder.

The sensitivity of this Adcock direction-finder was found to vary over the working range of wavelengths, owing apparently to an approach to a resonance condition in the transmission-line system. When used for bearing observations on vertically-polarized waves propagated horizontally, the mean sensitivity was of the same order as that of the simple loop set described above, i.e. bearings with a swing of 5° could be taken in a field intensity of about 100 μV per metre.

(3) ACCURACY OF DIRECTION-FINDERS

(a) Instrumental Accuracy

In order to obtain some idea of the instrumental accuracy of these direction-finding sets and the conditions to be fulfilled for a site on which they can be operated, simple calibration tests were first made on a flat open field some 250 yards square. A portable transmitter was set up in the centre of this field, and the direction-finder was taken to various positions in circles of radii 30 to 60 yards, the radio bearing being directly compared with the visual bearing. Out of 64 observations made in this manner, 59 were less than 2° in error, two of the remaining five being 3° in error. These errors refer to the mean observation of a pair taken by rotation of the coil, the difference between direct and reciprocal bearings being usually less than 2°. This latter figure has been reduced to less than 1° in later models of these sets. It is thus considered to be fair to state that the instrumental accuracy of these direction-finding sets for observed bearings lies within ± 2°.

Experiments carried out on the same and other sites indicate that the above accuracy is maintained at distances up to 600 yards from the transmitter. The results obtained at greater distances are dealt with in a later Section.

(b) Effect of Wires and Pipes on Direction-Finding Bearings

In the course of the calibration tests described above it was found that a long water pipe buried at a depth of 1 ft. 6 in. produced no effect on the bearings observed on a direction-finder placed over it, the height of the centre of the loop being about 6 ft. above the ground. Under similar conditions lengths of copper wire up to 50 yards long laid on the ground produced no effect; only when the wires were raised to the level of the base of the loop, and placed a few inches from it, were errors in bearing detectable. In later experiments it has been found that placing the direction-finding set immediately beneath a number of overhead telephone wires along a road produced a change in bearing of 3° on a distant station.

When short lengths of wire insulated from earth were placed vertically near the direction-finder, it was found that the effect was greatest when the wire was half a wavelength long. In this condition, the presence of the wire at a distance of 4 yards produced errors in bearing ranging up to 11°, the error being greatest when the direction of the wire was at right angles to the line joining transmitter and receiver, and practically negligible when situated on this line. The effect of such a vertical half-wave aerial became negligible when the aerial was removed to a distance of 30 yards from the direction-finder.

(c) Effect of large Aerials

In one phase of the experimental work the opportunity occurred to examine the effect of an aerial system the dimensions of which were large compared with the wavelength on which the direction-finder was operated. The system comprised four vertical-cage aerials used as an Adcock direction-finder for medium and long waves. Four aerials were arranged at the corners of a square of 320 ft. side. Each aerial was 70 ft. high and was formed of an octagonal cage of wires with a side of about 4 ft. When the ultra-short-wave loop direction-finder was used at a distance of about 70 yards from the centre of this system, errors in bearing were experienced ranging up to 11° on a transmitter a few miles away. When all the aerials were removed this error was reduced to 3°, which was considered to be the error due to other obstacles—chiefly trees—surrounding the direction-finding site. A single cage aerial at a distance of about 100 yards from the direction-finder produced a maximum error of 5° on a transmitter about 20 miles away, while when the cage was replaced by a single vertical wire of the same height (70 ft.) the error was less than 1°.

These values of the errors are intended to indicate the order of the effects observed rather than to be taken as absolute values, for it was found that the actual values depended upon the frequency employed and the direction of the transmitting station. In the case of observations taken on the Teddington station, about 11 miles away, the bearing was found to vary with the frequency in a quasi-periodic manner, indicating the presence of an interference pattern in the received field arising from the superposition of re-radiation from the aerial system on the incoming field from the distant source. In a similar manner it was found that if the portable transmitter was moved in a straight line away from the direction-finder to distances up to 100 ft. the bearings varied in a periodic manner over an amplitude of 4° or 5°, owing to the same cause. Although an attempt was made to study this interference pattern systematically it was found not to be susceptible to a simple analysis, probably owing to the complicated conditions prevailing at the site and to the fact that the aerials were not the only obstacles affecting the observations.

(d) Effect of a single Tree on Direction-Finding Bearings

Similar experiments to the above were carried out with the direction-finding set located near a single tree about 25 ft. high and with a spread of its branches of the order of 12 ft. A portable transmitter was used for these tests at a distance of 30 yards on a wavelength of 9·5 metres. With the direction-finder placed at a distance of 4 yards from the tree, an error in bearing up to 10° was experienced, the maximum effect being as in the case of the vertical wire, when the direction of the tree from the direction-finder was at right angles to that of the transmitter. When the set was moved to a distance of 20 yards from the tree, the maximum error was reduced to 1·5°. It is evident that a separation of 30 yards from such a tree would be sufficient to produce a negligible effect. On the other hand, later experience has shown that a large mass of trees may produce an appreciable error in observing the bearing of a distant transmitter with the direction-finder placed some 20 to 100 yards away.

(e) Effect of Buildings

An experiment was next carried out in which a portable transmitter was set up on an open site and operated on a wavelength of 8·0 metres, while the direction-finder was moved in a straight line along a road with various laboratory buildings on either side. Up to a distance of

280 yards there was a clear air line between transmitter and direction-finding set, but errors varying from − 9° to + 29° were experienced, accompanied by the effects of flat signal minima. In other experiments in similar "built-up" areas, where the direction-finding set was not in direct visual range of the transmitter, errors of ± 24° were experienced at ranges of 360 yards. It was usually found that, in these situations, the error of the observed bearings altered appreciably when the direction-finder was moved a few yards. All these effects are to be expected as the result of the interference of two or more waves arriving at the receiver after reflection from the surrounding obstacles.

(f) Conclusions as to Effect of Site Conditions and the Calibration of a Direction-Finder

The experiments described above provide considerable information as to the conditions to be satisfied for a site for an ultra-short-wave direction-finder in order that the observed bearings shall be free from the disturbing effects of obstacles in its locality. It appears that a flat, clear piece of ground some 100 yards in radius forms a satisfactory site, since the direction-finder is then practically unaffected by the presence of wires, trees, and similar obstacles. It has been shown, however, that a similar clear site is necessary for locating the transmitter if the "source error" which results from the re-radiation of waves from trees and other obstacles near the transmitter is to be avoided. Thus, in order to ascertain the actual instrumental accuracy of the direction-finder, it is necessary that a short-range calibration should be made on a site which is clear over a radius sufficiently large to accommodate both transmitter and receiver. The results obtained at three such sites will serve to illustrate the accuracy obtainable with the apparatus referred to in this paper.

(i) On a sports ground at Teddington, some 250 yards square, the errors observed in a complete calibration on a portable transmitter at a radius of about 60 yards were in general not more than 2°.

(ii) On a flat site at Dorney Common, Bucks, observations on a transmitter moved in a straight line away from the direction-finding set to distances up to 400 yards showed a maximum error of 2°. A similar run on another site has been carried out to the same order of accuracy at distances up to 650 yards.

(iii) In this case the direction-finding set was placed at the end of a concrete jetty, 1 120 yards long, in the middle of Queen Mary Reservoir, Staines, surrounded, except for the jetty, by an expanse of water of from 570 yards to 1 580 yards. When the transmitter was actually on the edge of the reservoir, the errors were very small (1° or less), except in positions near a wire fence, in which cases some source error was to be expected. Observations made over a distance of 360 yards at the projecting end of the jetty showed that the variations in the error were within ± $1\frac{1}{2}$°, so that the concrete itself had only a small effect on the bearings.

The results of the experiments described in this Section thus confirm the previous conclusion that the instrumental accuracy of the direction-finders employed is of the order 1° or 2°, and that for short ranges up to 1 mile, when both the transmitter and direction-finding set are on good sites and there is clear, level ground in between, the accuracy of observed bearings is generally within 2°.

(4) RESULTS OF DIRECTION-FINDING OBSERVATIONS AT GREATER RANGES (UP TO 25 MILES)

In order to ascertain the practicability of direction-finding on ultra-short wavelengths, a systematic series of tests was carried out in which the loop direction-finder was used at various ranges up to 22 miles from Teddington and in different directions.

(a) Local Observations on Teddington Transmitter

The fact that conditions between transmitter and receiver can seriously affect the accuracy of the observed bearings was first illustrated by some experiments carried out in Bushy Park at ranges up to about $\frac{3}{4}$ mile. The bearings observed on the main transmitter, situated at from 300 to 1 250 yards, were in error by amounts ranging up to 25°. It is considered that these errors were due to the effects of the large number of trees situated in the path of transmission, but to track down this cause in a satisfactory manner presents a somewhat difficult problem.

(b) Distant Observations on Teddington Transmitter

Table 1 gives the results of bearing observations made on the transmitter at the National Physical Laboratory, Teddington, operating on a wavelength of about 9 metres. Every endeavour was made to place the direction-finding set on a site which was considered to be good as judged by inspection, although no actual calibrations were made. In most cases the bearing was the mean of a number of consecutive observations, and in one or two cases repeat observations were made on a different day. In some instances these repeat observations were made with a different direction-finding set, and the results obtained were consistent. No difficulty was experienced in taking the bearings to an observational accuracy of 2° or less; for example, at the maximum distance of 22 miles the swing on which the observation was made was about 8° in some cases, the signal being inaudible at the minimum position. In some cases, at the shorter ranges, and particularly when the bearing was in error, larger swings were necessary and the minima were not always silent.

During the course of these tests it was noted that aeroplanes flying in the neighbourhood of the direction-finder produced considerable fluctuations of signal strength which sometimes rendered observations difficult.

An inspection of the results given in Table 1 shows that at 28 of the 43 sites tested the observed bearing was within the accuracy of ± 2° assigned to the instrument in Sections (2) and (3) above. In 15 cases the error was greater than 2°, the maximum recorded being 8°. It is to be noticed that there is a definite tendency for the larger errors to be of lesser magnitude as the distance of the direction-finder from the transmitter is increased, as was the case in the earlier work of 1931. While the conditions surrounding the transmitter, such as large masses of trees, and also buildings, may be a contributory influence in

Table 1

RESULTS OF OBSERVATIONS ON TEDDINGTON WITH LOOP DIRECTION-FINDER

($\lambda = 9$ metres approx.)

Site		From transmitter		Error in observed bearing
Ref. No.	Place	Distance	True bearing	
		miles	deg.	deg.
43	Hampton Court ..	1·0	14	−1
44	Sunbury Court ..	2·5	77	+1
45	Bushy Park ..	0·75	114	−4
46	Bushy Park ..	0·5	120	+8
47	Bushy Park ..	0·75	285	+8
39	Claygate	4·75	3	−6
40	Hersham	4·5	36	+3
41	Lower Halliford ..	4·75	77	+4
—	Queen Mary Reservoir	5·25	84	−1·5
31	Bedfont	4·25	112	0
21	Hatton	4·5	132	0
32	Heston	5·25	157	+2
33	Osterley	5·75	173	−2
35	Richmond Park ..	3·75	225	−6
36	Wimbledon Common	4·5	250	+1
38	Ewell Court ..	4·75	324	−3
42	Ruxley Farm ..	4·75	329	0
17	Effingham	9·7	20	−2
16	Ockham	9·7	32	−6
15	Pyrford	10·2	46	−1
14	Addlestone ..	8·0	67	0
13	Virginia Water ..	9·0	87	+1
12	Englefield Green ..	10·2	95	−4
11	Runnymede ..	10·0	101	−4
19	Sunnymeads ..	10·2	112	0
10	Dorney Common ..	14·2	113	−4
20	Datchet golf course	11·7	114	−4
26	Langley	10·5	124	0
24	Hillingdon	9·7	151	−1
23	Northolt	9·7	169	−3
22	Wembley	10·0	186	0
30	Guildford By-pass ..	21·0	40	0
28	Hog's Back ..	20·0	47	+1
27	Frimley	17·5	66	−1
7	Blackwater Flats ..	22·0	73	0
8	Caesar's Camp ..	18·0	83	−1
9	Winkfield	15·2	96	+1
1	Whitebrook Common	18·0	120	−3
2	Wooburn Common	18·2	127	0
3	Tyler's Green ..	21·5	135	+2
4	Coleshill	20·2	144	+2
5	Bovingdon ..	21·7	159	+1
6	Bedmond	21·2	171	−1

producing these errors, it is considered that at distances of more than 5 miles such direct influence is likely to be small. It would therefore appear that some portion of the errors, particularly at the greater distances, is due to deviations of the waves in travelling between transmitter and receiver. The arrival of two or more sets of waves at the receiver, and their consequent interference, is probably partly, if not wholly, the basis of production of this path error. In some cases where an appreciable error was experienced, it was found that moving the direction-finding set a few yards produced a change in the error; but the absence of such an interference pattern is not an infallible criterion for the absence of error. For example, Dorney Common (Site No. 10) is a wide expanse of open, flat common, and small changes in position (about 10 yards) of the direction-finder produced no effect on the observed bearing.

(c) Observations on Heston

During the course of the field observation work from which Table 1 was constructed, bearings were taken on the homing beacon at Heston Airport on a wavelength of 9 metres at such times as this station was found to be in operation. In this case the transmitter operates on a directional aerial system (Lorenz type) used for the guiding of aircraft towards the Heston airport at times of bad visibility. It employs two broad beams, on one of which a series of dashes is emitted while on the other is a corresponding series of dots. In the position midway between the beams, which in this instance is in the direction 84°, the two signals merge into a continuous dash. The results of the direction-finding observations obtained on this transmitter are shown in Table 2.

In making these observations it was generally found that both the signal strength and the signal minima were of good quality, although in some cases it was observed that the dot-and-dash signals did not pass through a minimum value at the same position of the direction-finding coil. As a result a bearing differing by not more than 2° was obtainable on the two series of signals. This effect is attributed to the fact that the two beams are reflected or scattered by different reflecting objects in their paths, so that the resulting waves arriving at the direction-finder produce different interference effects for the two sets of signals. This observation, in fact, tends to confirm the deduction that some of the errors in bearing experienced are the result of a path error, introduced at a position intermediate between transmitter and receiver.

(d) Observations on Alexandra Palace and other Transmitters over Extended Periods

With the regular transmissions available from the London television station at Alexandra Palace, the opportunity was taken to make regular observations at the working programme periods for several weeks. This programme was started in the winter of 1936 and continued in the summer of 1937. For this purpose two 1-hour periods were available, one each afternoon and evening in the winter, and an additional morning period in the summer. A continuous series of observations of bearings was made at the rate of about one bearing a minute, successive periods of 15 minutes being devoted to the sound and vision transmissions alternately. The direction-finder was set up in a hut at the site where the experiments with large aerials, described in Section 2(*c*) above, were carried out. It was known that this site was not perfect even with the aerials removed, and that a permanent error would probably be experienced. It transpired that this error was different for the two transmissions (sound and vision), due to the difference in wavelength. The observations carried out in the above manner showed that during any period of 15 minutes the bearings were in general constant to within 1°, but that occasionally a change of the bearing somewhat greater than this was observed from one period to a later one. A summary of all the results obtained is given in Table 3.

Table 2

RESULTS OF DIRECTION-FINDING OBSERVATIONS ON HESTON, WITH LOOP DIRECTION-FINDER

($\lambda = 9$ metres)

Site		From transmitter		Error in observed bearing	Error in bearing of Teddington (from Table 1)
Ref. No.	Place	Distance	True bearing		
		miles	deg.	deg.	deg.
41	Lower Halliford	6·25	19	−1	+4
31	Bedfont	3·5	21	+1	0
24	Hillingdon	4·5	152	−6	−1
22	Wembley	6·25	215	0	0
33	Osterley	2·25	244	0	−2
36	Wimbledon Common	7·75	294	−2	+1
—	Queen Mary Reservoir	5·75	24	+6	−1·5
42	Ruxley Farm	10·0	330	0	0
39	Claygate	9·75	345	+7	−6
3	Tyler's Green	16·5	129	0	+2
6	Bedmond	21·2	177	−1	−1

Table 3

SUMMARY OF OBSERVATIONS TAKEN ON TRANSMISSIONS FROM ALEXANDRA PALACE (SOUND AND VISION TRANSMITTERS)

Distance 20 miles, true bearing 67·5°

Date	Transmitter	Wavelength	Observed bearings		Deviations from mean bearing	
			Number	Error of mean	Extreme	Within 2°
		metres		deg.	deg.	%
Nov.–Dec., 1936	Sound	7·25	314	+0·8	$3\frac{1}{2}$	96
	Vision	6·7	432	−2·9	$1\frac{1}{2}$	100
July–Aug., 1937	Sound	7·25	122	−1·5	2	97
	Vision	6·7	144	+2·8	$1\frac{1}{2}$	100

In comparing the results for summer and winter it should be borne in mind that the large Adcock aerial system already described was removed in the meantime. These results were obtained by three observers, but no consistent difference was noticed indicating any systematic personal error. It will be seen from the Table that out of several hundred observations the maximum departure from the mean bearing was $3\frac{1}{2}$°, while 97·5 % of all the observed bearings differed from the mean by less than 2°.

It is to be concluded from these results that bearings taken on transmissions from the Alexandra Palace at a range of 20 miles show no diurnal variation exceeding about 2°. No estimate of seasonal effects can be made, owing to the change in receiver site conditions described above.

Similar observations were taken during the summer on transmissions from a television transmitter at Kensington, working on wavelengths of 6·39 and 6·12 metres respectively (frequencies 47 and 49 Mc./sec.). The results obtained from this source were of the same character as those described above.

During the course of this work, various observations were made on transmissions from American stations working on wavelengths between 8·5 and 10·7 metres (frequencies 25·6 to 28 Mc./sec.). In this case the observed bearings sometimes varied as much as 15° in the course of a few minutes' observations, although in most cases the variations were not greater than 10°. No opportunity has yet occurred to make a systematic investigation of this matter.

(e) Effect of Frequency on Observed Bearings.

During the course of the investigations it was noticed that in some cases the observed bearing was considerably in error although the site of the direction-finding set appeared to be satisfactory, while in other cases the observed bearing altered appreciably when the direction-finding set was moved only a few yards. The observations indicate the possibility of two or more waves arriving at the receiver and producing interference effects. In order to ascertain the validity of this deduction, some experiments were carried out in which the frequency of the transmitted waves was varied over a limited range while observations were made of the corresponding direction-finder bearings. The results of such experiments are given in Table 4.

Table 4

EFFECT OF VARYING FREQUENCY OF TRANSMITTER ON ERROR IN OBSERVED BEARINGS

Site No.	Transmitter	Site of direction-finder	From transmitter		Error in observed bearing on wavelengths of:—				
			Distance	True bearing	10·5 m.	9·7 m.	9·4 m.	9·1 m.	8·8 m.
					deg.	deg.	deg.	deg.	deg.
1	Portable	Dorney Common . .	100 yards	Various	—	+1·3	+0·5	—	+0·5
2	Portable	Dorney Common . .	300 yards	249°	—	−1·5	−1	—	−1·5
3	Teddington . .	Dorney Common (open site)	14·5 miles	113°	−5	−6	−6	—	−4
4	Teddington . .	Dorney Common (under telephone wires)	14·5 miles	113°	−3	−3	−3	—	−2
5	Teddington . .	Dorney Common (near trees)	14·5 miles	113°	−3	−5	−10	—	−10
6	Portable . .	Queen Mary Reservoir	0·5 mile	Various	−1	—	−0·5	0	+0·5
7	Teddington (Main aerials up)	Queen Mary Reservoir	5·25 miles	—	0	−2	−1·5	—	−2
8	Teddington (Main aerials down)	Queen Mary Reservoir	5·25 miles	—	+2·5	+2·5	+2·5	—	+1·5

The observations shown in this Table were taken at two sites which appeared to be good as the result of the calibration with a portable transmitter at a radius of 50 yards or more. The first of these, Dorney Common, was clear of trees and other obstructions for a radius of at least 200 yards, and in one direction the site was clear to a distance of over 400 yards. In this direction Test 2 shows that the error in bearing on the local transmitter was 1·5° or less, and that it did not change appreciably with frequency. With the direction-finder in the same position the error on the Teddington transmissions was between − 4° and − 6°. As a result of moving the direction-finding set very nearly under some telephone wires, the error decreased slightly (Test 4), while when the direction-finding set was moved to within 20 yards of a mass of trees the error ranged from − 3° to − 10°, according to the frequency.

At the much superior direction-finding site at Queen Mary Reservoir, Test 9 shows that on a local calibration at a radius of about $\frac{1}{2}$ mile across the water, the maximum error was 1° and did not vary with frequency within the errors of observation. On the transmissions from Teddington there was only a small variation with frequency, but it is seen that the error changed by some 3° or 4° when the main transmitting aerial at Teddington was lowered to the ground. This main aerial is of "T" type, 450 ft. long and 100 ft. high, and normally the ultra-short-wave transmitting aerial is slung about 50 ft. beneath the centre of this. There are also two other small aerials which were lowered for the above experiments. The angle subtended by the large aerial at the direction-finder at the reservoir is much less than 1°, so that it is not easy to understand the manner in which the aerial is directly responsible for the above change of 3° or 4°. It is possible that the position of the main aerial affects the distribution of radiation, and so indirectly is responsible for a change in the direction of some portion of the radiation arriving at the receiver. The results obtained on observations on the Heston transmitter at different places also indicated that the observed bearings were in some way affected by the direction in which the maximum radiation is projected from the transmitter.

These experiences indicated that conditions at the transmitting aerial could materially affect direction-finding bearings even at ranges of several miles, and some further experiments were conducted on this point as described in the next Section.

(5) STUDY OF EFFECT OF WAVE POLARIZATION ON DIRECTION-FINDER

(a) Effect of Horizontally Polarized Waves on a Loop Direction-Finder

In all the experiments described above, the transmitting aerial was vertical and thus the results obtained refer in all cases to the emission of vertically polarized waves. With a view to studying the effect of horizontal polarization, the aerial at the Teddington transmitter was supported in a horizontal position at a height of about 50 ft. above the ground. This aerial was connected to the transmitter by a tuned parallel-wire feeder as used in the previous experiments. The loop direction-finder was set up at the Queen Mary Reservoir site, where it was known that the error in bearing on Teddington, when using a vertical aerial, was 2·5° or less (Table 4). With the horizontal aerial the errors in bearings were in general greater than this; also they were subject to marked variation with orientation of the aerial and with alteration of the wavelength employed. The detailed results of these observations are given in Table 5, from which it will be seen that the errors observed range up to nearly 30°.

Table 5

EFFECT OF ORIENTATION OF HORIZONTAL PORTION OF TRANSMITTING AERIAL ON OBSERVED BEARINGS WITH LOOP DIRECTION-FINDER

Transmitter at Teddington; direction-finder at Queen Mary Reservoir; distance 5·25 miles; true bearing 84°.

Direction of top of transmitting aerial	Error in bearing for nominal wavelengths of		
	9·8 m.	9·2 m.	8·8 m.
deg.	deg.	deg.	deg.
17	+17	+11·5	+13
63	+ 5	+ 3	+ 6
82	+ 3·5	− 1·5	+ 1
119	0	+11	− 5
143	+17	+20	− 6
197	+10	+15·5	−28·5
243	+ 7	+ 7	+16
262	+ 3	+ 2	+11
299	−29	−11·5	− 7
323	−16	−13	+14

In general the errors are reduced to a minimum value when the aerial is in the line from transmitter to receiver, although this is sometimes masked by the large change in bearing with alteration in azimuth and wavelength. When the horizontal aerial was used, the signals were of adequate strength for direction-finding purposes, although the quality of the signal minima varied with the actual orientation of the transmitting aerial. In certain positions of the latter the bearing changed rapidly with orientation, thus sometimes making an observation difficult when the transmitting aerial was swinging slightly in the wind.

Tests made with the aerial disconnected showed that the vertical feeder was radiating to some extent, so that throughout the experiments a certain amount of vertically polarized radiation was superimposed on the horizontally polarized waves. For this reason the actual errors observed were to some extent dependent upon the adjustments of the feeder, and somewhat larger errors, up to 50°, were experienced at the same receiving site when emission of vertically polarized radiation was minimized by correctly matching, instead of using the feeder in the tuned condition.

Similar large errors in bearing on the Teddington transmitter were obtained at the Radio Research Station, Slough. In order to see whether the bearing changed with time for constant conditions at the transmitting end, observations were carried out during day and evening periods on five separate days spread over two weeks. The transmitting aerial was maintained in a fixed position, with the horizontal portion at approximately 90° to the direction of the receiver. Under these conditions the bearings were very flat; the mean error in the observed bearing was 78° and this was subject to a variation from one period to the next of as much as 21° on the same day. Although the transmitter was operating on the nominal wavelength of 9·3 metres throughout, the change in bearing may have been partially due to change in frequency of the transmitter, as special tests showed that changing the frequency by about 80 kc./sec. altered the observed bearing by some 10°. Furthermore, although the horizontal aerial was held during the tests as rigidly as circumstances permitted, some rotation of the aerial may have occurred and this would cause a change in error. It seems, however, that these variable conditions at the transmitting end cannot entirely account for the variations observed.

(b) Experiments with a Spaced-Aerial Direction-Finder

The effects described in the previous Section may have been due solely to the action of horizontally polarized waves on the loop direction-finder employed. It must be noted, however, that similar results would be obtained if the waves arriving at the receiver had a vertically polarized component, and if the direction of arrival showed a considerable lateral deviation from the line joining transmitter and receiver. It was in order to decide between these possibilities that the Adcock spaced-aerial direction-finder described in Section 2(*b*) was developed, since this instrument is insensitive to the action of horizontally polarized waves. When this instrument was completed, therefore, further experiments were carried out at the Queen Mary Reservoir site, in which a direct comparison was made of the bearings observed on both the loop and spaced-aerial direction-finder, while the horizontal aerial at the Teddington transmitter was rotated to various positions. The result of one of these tests made on a wavelength of 8·8 metres (frequency 34 Mc./sec.) is shown in Fig. 4. It is seen that while an error of nearly 60° was obtained with the loop when the transmitting aerial was broadside on to the receiver, the maximum error experienced on the Adcock set was only 6°. On a wavelength of 9·8 metres (frequency 30·5 Mc./sec.) the bearings with the Adcock set showed a total variation of about 4°, while the loop set gave errors up to about 30°.

It would appear from these experiments that although there is a vertically polarized component in the waves arriving at the receiver, this portion of the waves is not subject to any great lateral deviation. Such a vertically polarized component may have been present in the radiation leaving the transmitter, or it may have been produced indirectly by re-radiation from a secondary source. Although neither the above nor any other experiments so far conducted give proof that the horizontally polarized waves were confined to the vertical plane through transmitter and receiver, it seems reasonable to presume that these waves were subject to no greater lateral deviation than the vertically polarized waves. The large errors described in the previous Section must therefore apparently be explained in terms of horizontally polarized waves, the direction of travel of which is confined to the great-circle plane through transmitter and receiver.

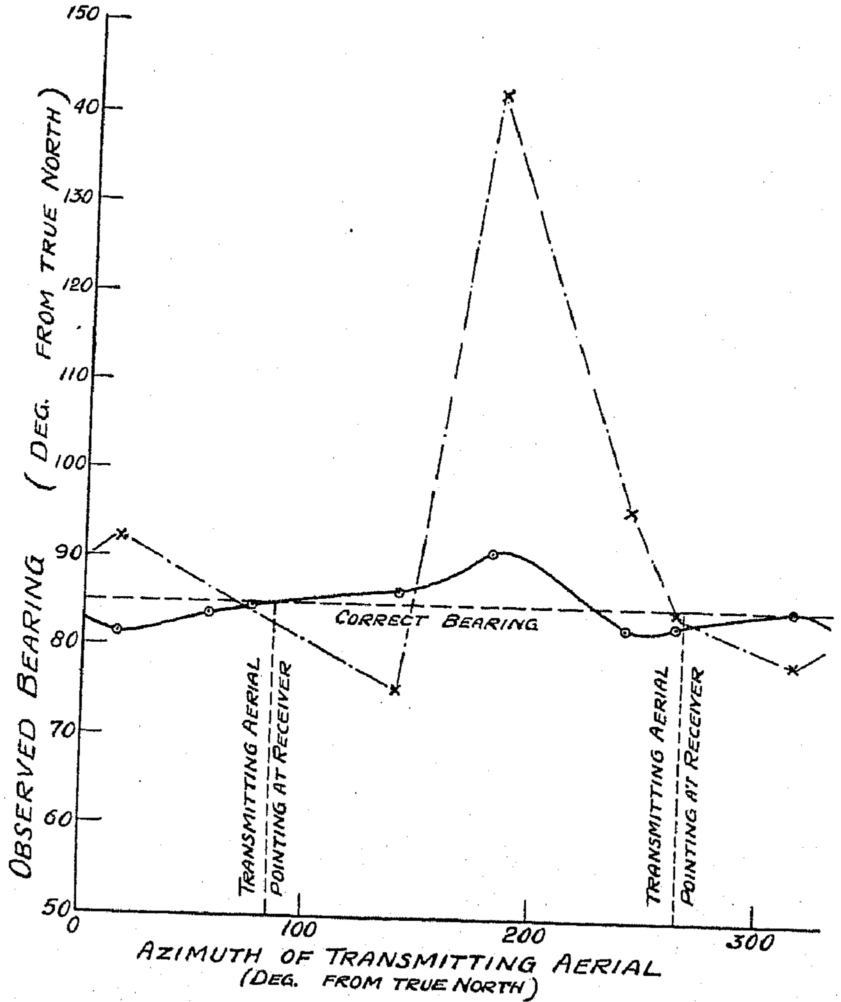

Fig. 4.—Comparison of the bearings obtained at Queen Mary Reservoir with the loop and Adcock receivers respectively, on the emissions from a matched horizontal dipole at Teddington. Working wavelength 8·8 metres (34 Mc./sec.) approx.

×- — - —× Loop bearings.
⊙———⊙ Adcock bearings.

(c) Further Investigation of the Action of the Loop Direction-Finder with Horizontally Polarized Waves

In seeking an explanation of the abnormally large errors experienced with the loop direction-finder, consideration was first given to the possibility of the loop receiver being susceptible to the action of horizontally polarized waves propagated horizontally. Such an effect might result, for example, from a serious lack of symmetry of the loop about a horizontal axis, so that the electromotive forces induced in the top and bottom sides of the loop produced a residual signal component in the receiver. In order to balance out such a component it would be necessary to rotate the loop away from the normal minimum position until an equal and opposite component was obtained from the e.m.f.'s in the vertical sides of the loop. The following tests appear to prove that the loop direction-finder employed was free from such a characteristic.

In the first place, when loops of different sizes and shapes were used on the direction-finder the error remained sensibly unaltered. Thus it appears that the error is not dependent upon a critical condition of lack of horizontal symmetry of the loops. Secondly, experiments carried out with a portable transmitter placed on the reservoir site at a distance of $\frac{1}{4}$ mile showed no large errors in bearing when a horizontal transmitting aerial was rotated in a horizontal plane. Furthermore, the observed bearing did not change appreciably as the transmitter was raised from ground-level to heights up to about 10 ft.

Thus we see that with a local transmitter presumably emitting horizontally polarized waves there is no appreciable error in observed bearing, while with a similar transmitter at a distance of over 5 miles large errors occur. The absence of error in the local case may have been due either to the small value of the horizontally polarized field arriving at the receiver owing to ground attenuation, or to the fact that the loop is not susceptible to horizontally polarized waves propagated horizontally. If the first explanation is true, it is to be expected that the horizontally polarized field from the Teddington transmitter would also be small. The presence of such large errors in this case seems, therefore, to indicate the presence of horizontally polarized waves arriving at an appreciable angle to the horizontal after being deflected in some part of the atmosphere between the transmitter and receiver.

Such downcoming radiation is of low intensity, since its presence only produces a serious effect when steps are taken to exclude the emission of vertically polarized radiation at the transmitter. At the shorter distance of $\frac{1}{4}$ mile at which the local test with the portable transmitter was made, the downcoming radiation was negligible compared with the residual vertically-polarized waves propagated along the ground. Furthermore, the results given in Table 4 showed that no large variations in bearing were observed on transmissions from stations, such as Alexandra Palace, employing vertical aerials. It thus appears that if any of such radiation is deflected from atmospheric regions in which a rotation of the plane of polarization might be produced, such horizontal downcoming radiation is weak compared with the vertically polarized direct radiation from transmitter to receiver.

Adequate opportunity has not yet occurred to study this problem in more detail, but a simple attempt has been made to measure the angle of incidence by comparing the signal intensity received on a loop and on an aerial in the manner employed some years ago on medium waves.* Experiments of this nature have been carried out at Queen Mary Reservoir, using vertically polarized radiation from Teddington. While the results obtained tended to indicate the presence of a downcoming component, certain difficulties were encountered which do not permit of a numerical interpretation of the results, e.g. fluctuations of signal strength due to the presence of aeroplanes in the neighbourhood.

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* See Reference (4).

(6) CONCLUSIONS AND FURTHER WORK

The experimental investigations described in this paper show that it is a practical proposition to use simple radio direction-finders with a rotating aerial system for wavelengths between 6 and 10 metres. While care is needed in the design and construction of the instruments to secure electrical symmetry and stability of operation, no difficulties are encountered other than those normally associated with ultra-short-wave technique. The rotating-loop pattern forms a compact, easily-transportable instrument suitable for much field investigation work. An inherent instrumental accuracy of such a direction-finder of from 1° to 2° can be obtained, but some difficulty is experienced in finding a suitable site on the ground on which such an accuracy can be utilized. Obstacles such as buildings, trees, and aerials, within a distance of about 100 yards of the direction-finder produce appreciable errors.

When a loop direction-finder is used on the ground to observe the bearing of ground transmitting stations using vertical aerial systems at distances up to 22 miles, the errors may be as great as 8°, although in the majority of cases the errors are less than 2°. The changes in bearing from day to day do not exceed about 2°. The direction-finders developed have been shown to be capable, under certain conditions, of giving bearings on commercial transmitting stations some 3 000 miles away. In this case the observed bearings are subjected to much larger variations than those given above.

When a horizontal aerial is employed at the transmitting station, serious errors are experienced with the loop direction-finder except at very short distances from the transmitter. These errors are attributed to the reception of horizontally-polarized waves, and they may be avoided by using a spaced-aerial or Adcock direction-finder, an experimental model of which is described in the paper.

The future work on this subject will be devoted to the development of an improved and more sensitive form of this Adcock direction-finder and to an investigation of its performance under a wide variety of conditions.

(7) ACKNOWLEDGMENTS

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THE MEASUREMENT OF THE LATERAL DEVIATION OF RADIO WAVES BY MEANS OF A SPACED-LOOP DIRECTION-FINDER*

By R. H. BARFIELD, M.Sc., Associate Member, and W. ROSS, M.A.

[From the National Physical Laboratory.]

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SUMMARY

The paper gives an account of some systematic measurements of the lateral deviations of wireless waves received at Slough from various short-wave transmitters over a period of several months. The measurements were made by means of a spaced-aerial direction-finder of the four fixed-loop type, and the first part of the paper is devoted to a description of the apparatus, together with an account of its performance in respect of instrumental accuracy, pick-up factor, and other characteristics.

The second part of the paper describes the measurements themselves, which were made photographically from a cathode-ray oscillographic goniometer. Some of the observations were made on pulse transmissions from Nauen ($\lambda = 30$ m.) and Dorchester ($\lambda = 37{\cdot}3$ m.), and the remainder on continuous-wave transmissions from Zeesen ($\lambda = 31{\cdot}4$ m.), Prague ($\lambda = 44{\cdot}6$ m.), and various American stations.

The results obtained show that the variations in bearings which arise from instrumental causes do not exceed about 2°, and are usually of the order of 1°. The records demonstrate that lateral deviation occurs for the above cases to an extent which depends on the range and type of the reflected waves observed. Deviations of 10° to 20° were recorded from the more distant stations, while in the case of Dorchester, distant 160 kilometres, the deviations were as much as 50° on occasions. From a study of these observations it is concluded that the effective points of reflection at the ionosphere may be as much as 50–100 km. out of the great-circle path.

(1) INTRODUCTION

The measurement of the azimuthal departure of radio waves from the great-circle path has been one of the main objectives of the direction-finding research at the Radio Research Station, Slough, for many years past, for it is clear that the extent and accuracy of such knowledge can alone determine to within what limits of accuracy it is, or will be, possible to carry out practical direction-finding. The early stages of this investigation gave information on the subject which has been of much value. Owing, however, to the existence of residual instrumental error in the apparatus then used it was not possible, from these early experiments, to obtain any positive measurements of lateral deviation, although it was possible to assign a limiting value to its magnitude in many cases.

This work, which has been described in an earlier publication,† opened up the possibility of developing improved practical direction-finding apparatus, and, since without such apparatus further work on lateral deviation could not be carried out, it became desirable from two points of view to devote time to this development work.†

As this work has progressed, however, it has been becoming increasingly apparent that the degree of improvement achieved thereby in direction-finding design is not accompanied by a corresponding degree of immunity from error in practice.‡ This conclusion is supported by others who have developed the Adcock system. Eckersley,§ experimenting with short waves, demonstrated the existence of extensive scattering in the skip distance, coupled with the very large deviation of waves arriving from transmitters at distances comparable with the earth's semi-circumference or for which the great-circle path passes near the North Pole.

Further evidence of scattering has been obtained by Ratcliffe and Pawsey,‖ employing a method based on the simultaneous measurement of fading at two adjacent sites and being thus independent of direction-finding apparatus.

It has, therefore, now become well established that a finite amount of lateral deviation exists which sets a definite and final limit to the accuracy to which a direction-finder can ever attain.

This paper describes the first stages of an investigation to determine these limits, dealing first with the experimental method adopted, and secondly with an account of the results of the first series of tests made with the apparatus.

(2) THE SPACED-LOOP DIRECTION-FINDER

(a) General Arrangements

In order to carry out this research an apparatus of the direction-finding type was required which would possess an instrumental accuracy—particularly as regards polarization and other variable errors—considerably higher than that of any which had hitherto been made. It was also necessary that this apparatus should be capable of giving almost instantaneous bearings, preferably with automatic recording thereof, and finally that it should be capable of working on pulse transmissions. The last two considerations could only be satisfied by a cathode-ray photographic recording technique. As regards the first, it had been found from a series of experiments that the complete elimination of polarization error with any form of spaced vertical-aerial system was very difficult, owing to the fundamental asymmetry introduced by the position of the earth's surface relative to the feeders or transmission lines connecting the aerials

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* Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 98.

† See Bibliography, (1).

† See Bibliography, (2) and (3).

‡ *Ibid.*, (4).

§ *Ibid.*, (5), (6), and (7).

‖ *Ibid.*, (11) and (12).

to the central receiver. The substitution of loop aerials for vertical aerials, however, provides a simple solution to this problem of balancing, as can be appreciated by a consideration of Fig. 1, which contrasts the essential symmetry of a loop with respect to its feeder with the essential asymmetry of various forms of vertical aerials in this respect. The possibility of entirely screening by pipe enclosure a loop aerial from the effect of static coupling (or of accumulated charges), on the other hand, tends to render accurate balancing unnecessary, so that the combination of these two factors makes it easily possible to produce a spaced-loop arrangement with negligible polarization error.

There are certain disadvantages introduced by such a substitution which render unsuitable the application of the spaced-loop system to practical direction-finding. It is not relevant to the subject in hand to discuss these here, but their nature is such that they are not at all serious when the apparatus is used for research only.

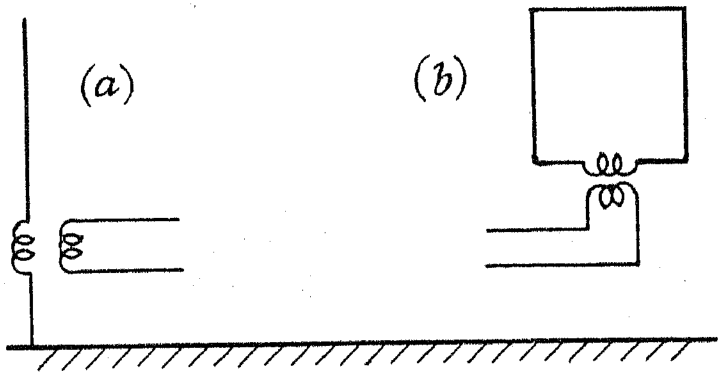

Fig. 1.—Asymmetrical and symmetrical coupling.

(*a*) Feeder coupled to aerial (asymmetry).
(*b*) Feeder coupled to loop (symmetry).

A spaced-loop direction-finder may be of the four-aerial fixed type or two-aerial rotating type. The latter type has been employed by Eckersley* for work of a similar nature, the results of which are referred to later in the paper, and it has also been developed at the Radio Research Station.† For the experiments here described, however, the four-loop system was employed and it is this which will now be described.

Four loops are arranged at the corners of a square of 12 metres diagonal, as shown in Fig. 2(*b*), and the diagonally opposite loops are coupled together through transformers and transmission lines so as to be in opposition. The system thus corresponds to an Adcock direction-finder in which loops take the place of vertical aerials. Each loop can be rotated about its own vertical axis, and all four are set to point in the same direction. Since the system is only used for downcoming waves of complex and variable polarization it will be clear that, as far as the "wanted" pick-up of the system is concerned, it does not matter which direction is chosen, provided each loop is accurately aligned thereon, for then the same field component will act on each loop and the output from each pair will depend solely on the phase differences. To lessen the risk of "unwanted" pick-up resulting from currents induced in the feeders, however, one of the two arrangements in which the loops all make angles of 45° with the feeders has been chosen for all observations made up to the present, as this arrangement is not only the most symmetrical but also the most suitable for the process of lining-up described below.

It is clear that a spaced-loop system, even when the loops are so oriented as to be acted upon wholly or partly by horizontally polarized waves, will be immune from polarization error, provided that the loops are accurately aligned in parallel planes and that the feeders connecting them do not pick up.

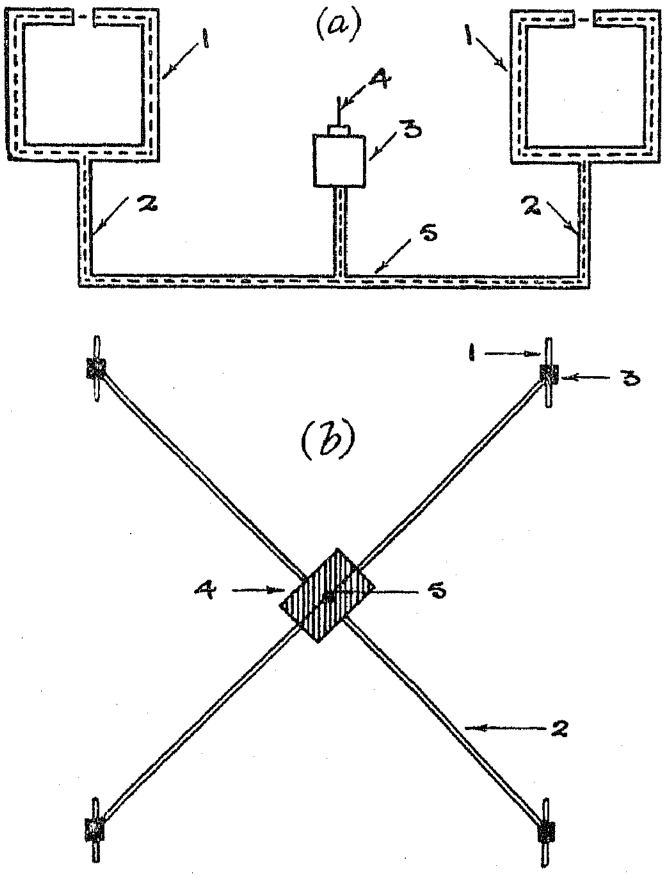

Fig. 2.—General arrangement.

(*a*) Elevation.

(1) Screened loop.
(2) Feeder.
(3) Central screening cage.
(4) Lining-up transmitter and vertical aerial.
(5) Feeder screen.

For simplicity the loops are here shown in line with feeders; actually they are at 45° as shown in (*b*).]

(*b*) Plan.

(1) Screened loop (1 m. square).
(2) Screened feeder.
(3) Screening box containing tuning condenser and matched transformers.
(4) Cage containing screened cathode-ray receiver.
(5) Central radiating aerial for lining-up.

(b) Constructional Details

Each loop consists of four separate turns of insulated wire which, by series-parallel switching, can be made to form from one to four turns. These wires are enclosed in a rigid tubular screen formed of $1\frac{1}{2}$ in. diameter aluminium piping jointed together at the corners and having an insulated gap at the centre of the top member. The wire is so arranged on spacers that it occupies approximately the central axis of the tube. The method of tuning the loops and of coupling to the feeders is shown in Fig. 3. The tuning condenser and transformer are housed in a screening box fixed at the centre of the frame, the leads from the loop being brought into this box through a metal pipe fixed at the centre of the bottom limb of the frame. The construction is illustrated in Fig. 4. It was ascertained by means of special tests that the pick-up factor of the loop was not reduced by the screening pipe surrounding it.

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*See Bibliography, (13).

† *Ibid.*, (14).

The transformers are wound on ribbed formers and are mounted inside a copper pot. The primary is in parallel with the loop, and the windings are so designed that the circuits are matched at wavelengths of 40 and 100 metres (frequencies of 7·5 and 3 Mc./sec.), which gives approximate matching over the wavelength range 30–100 metres. There is an electrostatic screen between the windings of the transformer. The screened twin feeders between the loops and the central hut are of ordinary twin lead-covered electric lighting cable. They are buried to a depth of a few inches, partly to bring them in close proximity to the surface of the earth and partly to protect them from mechanical damage, and are brought up along the outside of the central axis of the loop to the screening box, to which the screen is bonded.

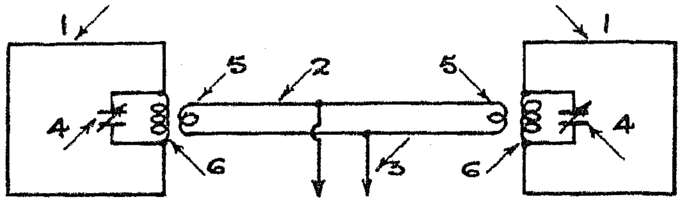

Fig. 3.—Circuit diagram.

(1) Loop.
(2) Feeder.
(3) Input to amplifier.
(4) Tuning condenser.
(5) Transformer secondary.
(6) Transformer primary.

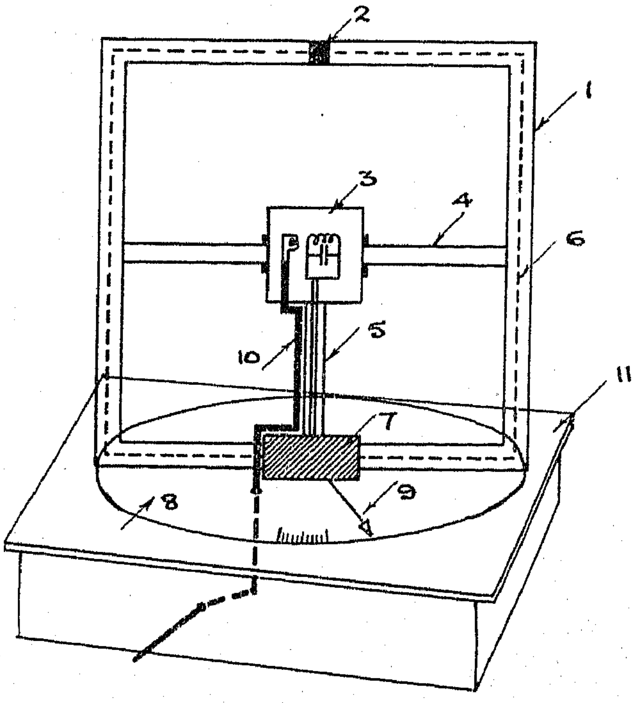

Fig. 4.—Loop details.

(1) Screening pipe containing loop and rotating about central vertical axis.
(2) Insulating gap in screen.
(3) Metal box containing tuning condenser and feeder transformer.
(4) Insulating supports.
(5) Metal tube containing leads from loop to (3).
(6) Four single loops.
(7) Series-parallel connection box.
(8) Scale.
(9) Pointer.
(10) Screened feeder to cathode-ray apparatus.
(11) Wooden supporting platform.

At the centre of the system inside the hut is fixed a screening cage approximately 6 ft. cube, consisting of a wooden framework covered with wire netting of 1 in. mesh to which the transmission lines are bonded at their point of entry at the centre point of its floor. The four feeders then pass into a small screening box (called the " mixing " box) where they are connected to a series of switches and variable resistances as shown in Fig. 5. By means of the switch the two incoming feeders constituting a pair are joined together (in conjunction or opposition as desired) and to the common input feeder leading to one of the amplifiers, while the resistance can be made to vary the relative output of the loops in order to obtain a perfect balance. This enables the process known as lining-up to be carried out in the manner described in another Section.

The two feeder-pairs from the " mixing " box pass directly to the input transformers of the pair of matched amplifiers whose outputs are connected to the plates of the cathode-ray tube. This combination of two amplifiers and cathode-ray tube constitutes a cathode-ray comparator. Essentially similar apparatus, the working thereof, and the methods of photographic recording adopted, have already been described in other publications.*

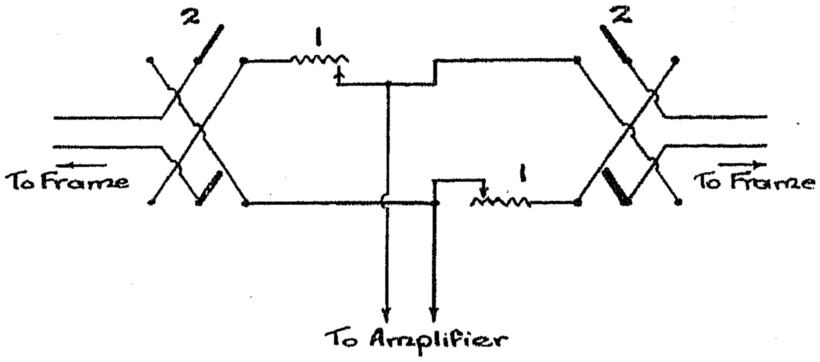

Fig. 5.—Circuit diagram of mixing box.

(1) Variable resistance.
(2) Reversing switch.

(c) Lining-up the System

It will be clear from the principle of action of the direction-finder that it must be assumed that, provided each loop is acted upon by an equal and co-phasal field, it must with its associated feeder contribute exactly co-phasal and equal e.m.f.'s to the central points of the system, and that on passing through the two amplifiers the signal from each pair is subjected to the same gain and phase-change. To attain this end all corresponding parts of the system are made as nearly equal as possible, but in practice it is not possible to attain in this way the degree of equality necessary to ensure the desired overall accuracy of the instrument, and certain controllable adjustments are provided for the final matching. To make these adjustments it is necessary to provide a local source of energy which, acting equally on all four aerials, enables the adjustments to be made which give equal output. This is provided by installing a transmitter with a short vertical aerial on the top of the screening cage at its geometric centre, and tuning it to radiate at the frequency of the signal to be received [see Fig. 2(*a*)]. With this switched on, fine tuning adjustments can be made in each loop, together with further resistance adjustments and the usual lining-up process, stage by stage, of the amplifiers. In this way the two loops of a pair can be balanced against one another to give the required degree of equality, and the two pairs can be equalized by appropriate adjustments to the amplifiers. The degree of equality required for a given accuracy, and details of how it is attained, are given in another Section.

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* See Bibliography, (15).

(d) Pick-up Factor

The " pick-up factor " of an aerial system is defined as the quotient of the potential difference across the first grid circuit of the amplifier and the field strength existing at the aerials. It is a measure of the sensitivity of the aerial system, and is determined by the size of the loops, the spacing of the aerials, the magnification factor of the tuned circuits, and the efficiency of the coupling transformers. It is found that the attenuation of the transmission lines themselves is negligible.

The pick-up factor of the system has been measured by experiment and found to have the values shown in Table 1, for a wave travelling along the ground.

Table 1

λ	Frequency	Pick-up factor
metres	Mc./sec.	metres
100	3·0	1·0
75	4·0	1·5
45	6·7	3·0
33	9·0	6·0

These values were found to be in close agreement with those calculated from the known dimensions and characteristics of the various parts of the aerial system. For frequencies above 4·0 Mc./sec. the loops were used with only one turn on each. It might appear that the pick-up factor on any frequency could be increased considerably by increasing the number of turns used on each loop. Theoretical considerations, however, have shown that this is not so, and experiment confirms these conclusions. Briefly, the reason for this is that while, by increasing the number of turns on any loop, the e.m.f. picked up by the loop is increased proportionally, the inductance of the loop and its high-frequency resistance are both increased much more rapidly, more or less proportionally with the square of the number of turns. This rapid increase of the resistance of the loop has exactly the effect of cancelling the increase in pick-up factor brought about by the increased e.m.f. received by the loop itself.

Two turns were used on the loops at the lower frequencies, chiefly to achieve a high inductance in the loop circuits and thus to facilitate tuning of the loops.

It should be borne in mind that since the pick-up factor of the system depends on the angular spacing of the aerials, it is a function of the angle of incidence of the wave, being in fact proportional to the sine of the angle of incidence. Thus the pick-up factor becomes much smaller for steeply downcoming waves.

(3) INSTRUMENTAL ERROR

Spaced-aerial direction-finders are liable to instrumental errors of several different types. It is proposed in this Section to deal with such errors as apply to the spaced-loop direction-finder, and to indicate the performance of the instrument in respect of such errors and to show to what extent the errors which remain in the instrument affect the accuracy of the results obtained.

(a) Polarization Errors

It is important to consider to what extent the spaced-loop direction-finder is liable to polarization errors, i.e. errors whose magnitude and sign depend on the state of polarization of the received waves. With this direction-finder it is found the polarization errors may arise from two causes.

Firstly, energy may be picked up by the outer screen of the horizontal feeders running between the aerials and the receiver, and transferred to the aerials by radiation or by electromagnetic coupling between the aerials and the outer screens of the feeders. It is clear that, in the presence of downcoming waves, if the loop aerials are so oriented as to receive energy either from the vertically polarized component of the wave only or partly from the vertically polarized component and partly from the horizontally polarized component, such transfer of energy from the screening cover of the transmission lines to the aerials would give rise to polarization errors, since the energy received by the screening cover of the lines is derived from the horizontally polarized component of the wave.

Secondly, if the planes of all four loops are not exactly parallel, polarization errors will again arise, since the loops will not be all equally receptive to the vertically or horizontally polarized component of the received waves.

Returning to the case of the first type of polarization error mentioned above, it is evident that the magnitude of the error involved will depend on several factors, viz. the strength of the horizontal electric field at the transmission lines, the " wanted " pick-up of the aerials themselves, and some factor in the nature of a coupling coefficient representing the ratio between the e.m.f.'s induced in the loops by currents flowing along the outer screen of the transmission lines and the e.m.f. picked up by that screen on account of the horizontal electric field of the wave. It may be noted that, in the special case in which the loops are so oriented as to be receptive only to the horizontally polarized component of the received waves, the error introduced is not strictly a polarization error, since both the " wanted " and " unwanted " pick-up are due to the same component of the wave. The error, however, does vary with angle of incidence in this case and is therefore still undesirable.

For simplicity of treatment, however, we shall deal with the case in which the planes of the loops are parallel to the vertical plane of arrival of the wave, so that the loops are receptive only to the vertically polarized component of the wave, and the " unwanted " pick-up is due to the pick-up of the horizontally polarized component by the outer screen of the transmission lines. The polarization error resulting when the planes of the loops make any other angle with direction of arrival of the wave can be shown to be of the same order of magnitude as for the case here considered. We shall further suppose that the wave arrives along the direction of the line joining one pair of aerials and use the following notation:—

A = area of one loop in square metres.
n = number of turns on loop.
d = distance between loops.
λ = wavelength.
E_V = intensity of vertically polarized component in wave.
E_H = intensity of horizontally polarized component in wave.
Y = intensity of horizontal electric field at the outer casing of the transmission lines.
K = number representing the "coupling coefficient" between outer screen of transmission lines and loops referred to above.
ρ_v = reflection coefficient of earth for vertically polarized waves.
ρ_h = reflection coefficient of earth for horizontally polarized waves.
e = error.
θ = angle of incidence of wave.
α, β = angular misalignment of one loop in horizontal and vertical planes.

We have the following:—

$$\text{"Wanted" pick-up} = E_V(1+\rho_v)\frac{2\pi An}{\lambda}\cdot\frac{2\pi d}{\lambda}\sin\theta$$

$$\text{"Unwanted" pick-up} = YdK.$$

Then the maximum value of e, obtained by assuming the "wanted" and the "unwanted" pick-up are in phase, is given by

$$\tan e = \frac{YK\lambda^2}{E_V(1+\rho_v)4\pi^2 An\sin\theta}$$

The coupling coefficient K has been measured experimentally in the following manner. The transmission lines from one pair of aerials were removed from their trenches under the ground and suspended close to the ground but not touching it. These transmission lines were each encircled by a toroidal inductance, the two toroids being fed from a common oscillator and so wound as to induce currents in the same direction in the two halves of the transmission line. Knowing the current in the toroids (measured by a thermo-junction) and the coefficient of mutual induction between the toroids and the lines, the e.m.f. induced in these lines can be calculated. The e.m.f. induced in the loops by the currents flowing in the lines is observed by noting the signal produced in the receiver, the two loops being connected together in the same sense as when used as part of the direction-finder. By injecting a known e.m.f. directly into the loops and observing the value of e.m.f. required to produce the same signal in the receiver as that produced by the currents induced in the outer screen of the transmission lines, the factor K can be found. The direct injection into the loops can be accomplished in several ways. One method is to insert a small resistance at the centre of the loop and to connect the output from a signal generator across the resistance. Another method, suitable in cases when it is not possible to reach the centre of the loops, is to radiate a signal from a small oscillator the field strength of which at the loop can be measured or calculated.

The value found for K, by experiment, was of the order of 10^{-4} for frequencies of 3 and 8 Mc./sec. This means that the error for a "standard wave" (angle of incidence 45° and equal vertically and horizontally polarized components) is not greater than 0·1° at these frequencies, and while errors of greater magnitude can occur they can only do so at the expense of signal strength. Thus errors of the order of 1° would occur only when the amplitude of the signal fell to about one-tenth of its normal value, that is to say when the "wanted" component in the wave became very much smaller than the "unwanted" component. As the state of polarization of ionospheric waves is constantly changing, provided bearings are only noted when the amplitude of signals is near the maximum the polarization error from this source is negligible.

As far as the second type of polarization error is concerned, namely that arising from the loops not being all strictly parallel, we may again consider a simple example to examine to what degree of accuracy the loops must be set up. The apparatus is supposed to be disposed in the same manner as described above, with the planes of the aerials parallel to the vertical plane of arrival of the wave and the line joining one pair of aerials pointing in the direction of arrival of the wave. We shall suppose that three of the loops are exactly parallel to the plane of arrival of the wave, while the fourth, which we shall assume to be one of the pair on the line at right angles to the direction of arrival of the wave, makes a small angle α with that direction.

The "wanted" pick-up for the pair of loops lying along the direction of arrival of the wave will be, as before,

$$E_V(1+\rho_v)\frac{2\pi An}{\lambda}\cdot\frac{2\pi d}{\lambda}\sin\theta$$

In the case of the other pair of loops, lying on a line perpendicular to the direction of arrival of the wave, the pick-up due to the vertically polarized component is zero, since it is equal (approximately) for the two loops which are in opposition, but since one loop makes a small angle α with the direction of arrival of the wave there is a residual "unwanted" pick-up due to the horizontally polarized component of amount

$$E_H(1+\rho_h)\cos\theta\frac{2\pi An\sin\alpha}{\lambda}$$

The maximum error, assuming the "wanted" and "unwanted" pick-up to be in phase, is given by

$$\tan e = \frac{E_H}{E_V}\cdot\frac{(1+\rho_h)}{(1+\rho_v)}\cdot\frac{\lambda}{2\pi d}\cot\theta\sin\alpha$$

For a standard wave $E_H = E_V$ and we may write $\rho_h = \rho_v = 1$ (approximately), from which it appears that e is of the same order of magnitude as α for a standard wave. Again, the error may be many times as big, but only when the ratio E_H/E_V becomes large or θ so small that the "wanted" pick-up becomes very small also. Thus, except for very small angles of incidence, if observations are confined to those made when the amplitude is large, the polarization errors due to this cause will be of the same order of magnitude as α. It can be shown that in the more complex case, in which the planes of the loops do not coincide with the plane of arrival of the wave and thus each receives signals from both vertically and horizontally polarized components, the result is broadly the same. It therefore appears that misalignment of any loop from the plane parallel to the others gives rise to polarization errors of the same order of magnitude as the angular misalignment, except when the signal strength falls to small dimensions on account of changes in the state of polarization of the wave.

What has been said above refers to azimuthal misalignments, but it is also true that if all the loops do not make equal angles with the vertical, polarization errors will arise. Calculations show that if one loop is out of the vertical by an angle β the standard-wave error is of the order of $\beta/|\sqrt{\kappa'}|$, where κ' is the complex dielectric constant of the ground.

In practice, it has been found that by making use of sights attached to the frames of the loops and training the loops on distant posts spaced apart the same distance as the loops, the error in angular alignment of the loops has been kept to a value less than 0·2°.

As an indication that polarization errors in this instrument are as small as the foregoing would appear to indicate, it may be noted that in the course of making observations on the direction of arrival of pulses from Dorchester (distant 160 km. from Slough) using a wavelength of 37·3 metres, occasions have occurred when the two components of the first-order F_2 echo, separated in time on account of magneto-ionic splitting, have been observed to give exactly the same bearing. These two components are elliptically polarized, having opposite senses of rotation of the ellipse in the two cases. Polarization errors would thus have opposite signs in the two cases.

Occasions have also been observed when the two components did not give the same bearing, but these appear to be genuine cases of difference in the direction of arrival of the two. If such difference in apparent direction were due to polarization error it would always exist. This, however, is not in agreement with observed facts.

(b) Other Forms of Instrumental Error

Of the other forms of instrumental error, the next in importance after polarization errors are those which arise from lack of balance between the aerials. Such errors will arise, for instance, if, when equal and co-phasal fields exist at the aerials, the signals produced at the receiver by all four aerials are not equal.

The theory of this type of error has been discussed in another paper,* and all that requires to be said here is that the magnitude of the error depends on the ratio of the signal obtained from a pair of aerials connected in opposition to the signal obtained from the pair connected in conjunction when equal and co-phasal fields exist at the two aerials of the pair. Now, as mentioned elsewhere in the paper, in order to achieve a sufficiently high value of pick-up factor the aerials used in this system were tuned to resonance. As stated in an earlier Section, it is found, in practice, that if the four aerials are separately tuned to resonance the phases and amplitudes of the currents in them are not in general sufficiently equal, and some methods of fine balancing are required. The method finally adopted makes use of a small oscillator with a short vertical aerial located exactly at the centre of the system. The four aerials are set up in such a manner that their planes are parallel and at 45° to the line joining their centres to the centre of the system. It is assumed that the oscillator will produce equal and co-phasal fields at all four aerials. After tuning all four aerials to resonance (i.e. until the amplitude of signal received from the central oscillator is a maximum) the opposite aerials in each pair are opposed. Ideally, the signal obtained from a pair of aerials should then be zero. In general it is found not to be so. By making slight adjustments to a trimming condenser connected in parallel with the main tuning condenser of one aerial of each pair and to a small variable resistance connected in series with the feeder from that aerial, the residual signal from a pair of loops connected in opposition can be controlled until its amplitude in comparison with the signal obtained either from one loop alone or from the pair connected so as to be in conjunction, is sufficiently small for the degree of accuracy required. Having " balanced " each pair of aerials in this way the aerials in each pair are connected so as to be in conjunction with respect to the signal from the central oscillator. They are then in opposition with respect to a distant oscillator, and are thus in the operating condition. With the members of each pair connected in conjunction, the signals induced in each pair ought to be equal and co-phasal, and these signals, after passing through a pair of identical amplifiers to the orthogonal pairs of deflecting plates of a cathode-ray oscillograph, ought to produce on the fluorescent screen of the oscillograph a line at 45° to the axes of deflection of the tube. The amplifiers themselves are balanced stage by stage as described in other publications,* so that when equal and co-phasal signals are applied to the input of each the trace on the oscillograph screen is a line at 45° to the axis.

It is found that when the two pairs of aerials are connected, one pair to each amplifier, with the aerials in conjunction with respect to the central oscillator, the figure on the oscillograph screen when signals are received from the oscillator is usually a thin ellipse whose major axis is nearly but not quite at 45° to the axis of the tube. This indicates that, while the members of each pair have been " balanced " with respect to each other, the one pair is not " balanced " with respect to the other pair. The necessary balance can, however, be easily obtained. All that it is necessary to do is to make a small adjustment to the tuning of one stage of one of the amplifiers or to the attenuator of the amplifier, or to both, until the figure on the tube takes the form of a line at 45° to the axis of deflection of the tube. The apparatus is then ready for use on the frequency at which the adjustments have been made.

It is necessary to deal with several questions of major importance in connection with instrumental errors arising from lack of balance or adjustment of the aerial systems. In the first place, the absolute accuracy of the apparatus after adjustment depends essentially on the accuracy with which the central oscillator produces equal and co-phasal fields at the four aerials. If it does not do this exactly, an error will be introduced by carrying out the aerial adjustments and balancing on the assumption that it does. Such an error will, however, always remain the same for a given direction unless changes are made in the disposition of the oscillator and aerials, since it is found in practice that the aerial system can be balanced to an accuracy better than ± 1°. There are some indications that the central oscillator does not produce exactly equal and co-phasal fields at the four aerials, since the mean of a large number of bearings taken on a distant fixed transmitting station usually shows an error of 2° or 3°. The means of small groups of observations on the same station taken at different times, however, usually lie within ± 1° of the mean of all observations.

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* See Bibliography, (16).

* See Bibliography, (15).

Once the aerial system has been balanced, it is important that balance should be maintained at least during the period that the apparatus is used. It has been found that in favourable circumstances (i.e. fine, dry, calm weather) the aerial balance remains sufficiently good for periods of as much as one or two hours continuously, and even in adverse circumstances the balance is nearly always maintained for at least 15 to 30 minutes continuously. In very damp weather, however, there is usually a steady drift in the balance, necessitating re-adjustment after such a period. In operation, the balance and general line-up of the apparatus is frequently tested, and readjustments are always made whenever a drift great enough to change the instrumental error by more than 1° has occurred. In addition to such slow changes in balance, there is usually also a more or less rapid " flutter " of the balance, particularly on very windy days. The " flutter " is never greater than would cause an oscillation of the bearing of any station by more than $\pm \frac{1}{2}°$, except for waves arriving at very small angles of incidence, say less than 30°. For smaller angles of incidence the uncertainty can be taken as inversely proportional to the angle of incidence.

The process of adjusting the apparatus for operation on any frequency has been described. It has been found by experiment that the adjustment of the apparatus only holds over a limited band of frequencies in the neighbourhood of the frequency at which adjustment was made. This appears to be essentially due to the fact that the aerials in operation are tuned to the incoming signal. Theoretically one might suppose that four identical aerials accurately matched to each other at one frequency would remain so matched. In fact, however, this state of affairs is unattainable, chiefly because the process of balancing does not adjust every part of one aerial to be identical with the corresponding part of the others, but simply ensures that the output of one aerial as a unit shall be identical with that of the others at the frequency of adjustment. It has been found that the frequency band over which the adjustments of the aerials hold is about 10 to 20 kc./sec. on each side of the frequency of adjustment. Outside these limits the error due to aerial unbalance becomes steadily greater than the instrumental accuracy of ± 1° aimed at. A further limit to the frequency range of the instrument for any particular adjustment is imposed by the limitations of the amplifiers available for use. It was found necessary to perform the " lining-up " process for these within about 3 to 5 kc./sec. of the frequency on which it was desired to operate. In practice, adjustment and lining-up of the apparatus is always carried out on as nearly as possible the exact frequency on which it is desired to operate, the central oscillator being tuned to produce a low audio note with the received signals if the latter take the form of continuous waves.

Finally, the apparatus is liable to other minor errors, such as errors arising in calibration of the protractor drawn on the cathode-ray tube face. This calibration is done at the intermediate frequency by the aid of a specially constructed potentiometer giving 10° steps round the protractor. Errors arising in calibration have a maximum of 2° and are generally considerably less. They are of course unimportant when studying changes in bearing and only affect the absolute accuracy of the instrument, since they are obviously fixed errors of the instrument.

It is convenient here to summarize very briefly the state of development of the apparatus so far as instrumental accuracy is concerned. The absolute accuracy of the apparatus is such that fixed errors in bearing of the order of 2° or 3° may be expected. Such errors will be constant for a given azimuth, though they may vary to some extent with angle of incidence since the effective spacing of the aerials depends on the angle of incidence. For a given ray, the angle of incidence of which remains within a few degrees of a fixed value, such errors in the absolute accuracy of the apparatus will not give rise to any variations in the apparent bearing. This is an important point since, in its application, the apparatus is used to investigate changes in bearing, the absolute accuracy of the bearings being of lesser importance.

In addition to such more or less fixed errors, there remain residual variable instrumental errors arising on account of fluctuations of the aerial balance, or very small fluctuations due to polarization error. These, however, would not give rise to fluctuations in bearing greater than ± 1° for all rays except such as are incident at small angles (less than, say, 30°) to the vertical. For such small angles of incidence the probable fluctuation in bearing is approximately inversely proportional to the angle of incidence.

Spurious fluctuations in bearing can arise in yet another way when receiving a continuous-wave signal which may be composed of several rays at different angles of incidence. If the instrumental error is different for the different rays each ray would have a different apparent bearing, and it can be shown that in these circumstances the apparent bearing of the composite continuous-wave signal will fluctuate as the relative phases of the separate rays change. The form of the trace on the cathode-ray tube will also pass through various forms of ellipse. At the same time the amplitude of the signal will vary and, at times when the amplitude is least, the apparent error of the bearing of the continuous-wave signal may be several times as great as the difference in the instrumental errors for the separate rays. Thus in the course of such fading the apparent bearing will fluctuate. Provided, however, note is only taken of the bearing at large amplitudes, say when the amplitude is at least as great as 0·5 times the maximum, such spurious fluctuations will not exceed ± 2° at the utmost.

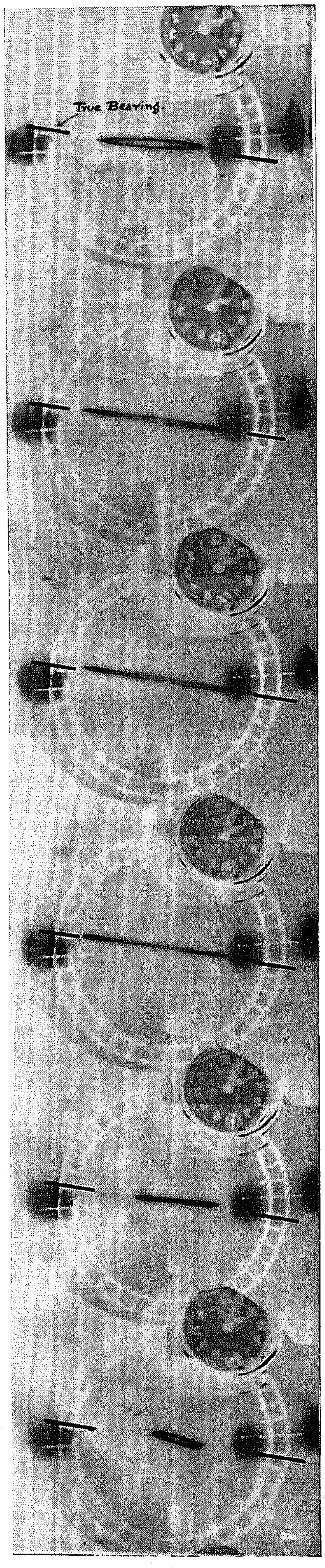

Fig. 7(a).—Zeesen (31·4 m.); 0955 G.M.T.; 2.10.36. Exposure = $\frac{1}{25}$ sec.; interval between photographs = 5 sec. Receiver beyond skip distance.

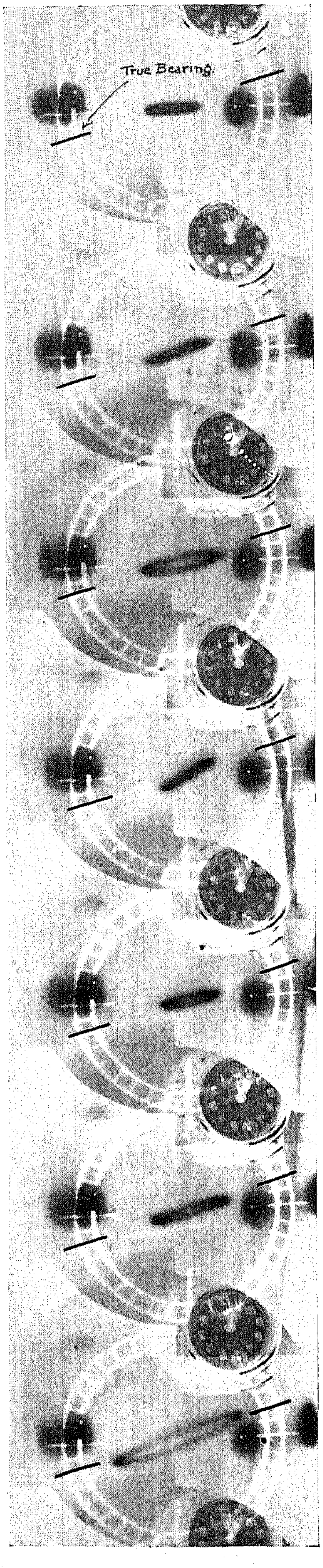

Fig. 7(b).—Sayville (36 m.); 2353 G.M.T.; 27.10.36. Exposure = $\frac{1}{25}$ sec.; interval between photographs = 2 sec. Typical bearings.

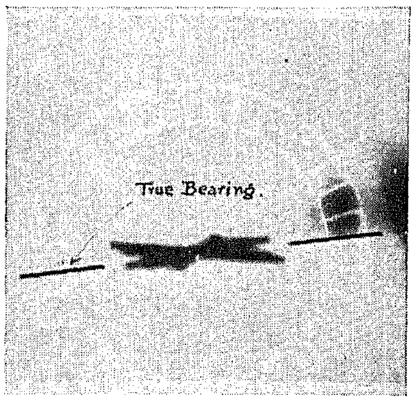

Fig. 7(c).—Prague (44·6 m.); 1530 G.M.T.; 8.3.37. Exposure = $\frac{1}{25}$ sec. High-speed morse transmission giving complex pattern.

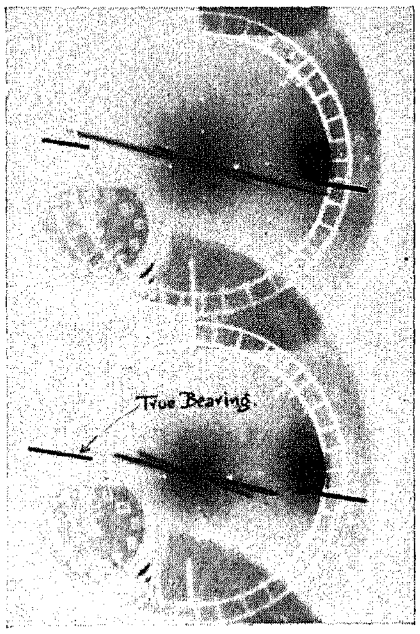

Fig. 7(d).—Nauen (30·6 m.); 1300 G.M.T.; 15.12.36. Exposure = $\frac{1}{10}$ sec. Pulse transmission, 1st- and 2nd-order F_2 rays giving different bearings.

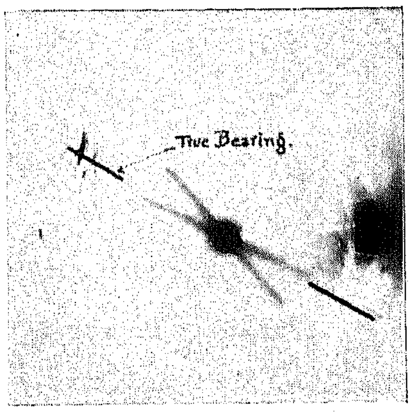

Fig. 7(e).—Dorchester (37·3 m.); 1200 G.M.T.; 7.5.37. Exposure = $\frac{1}{8}$ sec. Pulse transmission. Ordinary and extraordinary components of ray giving different bearings.

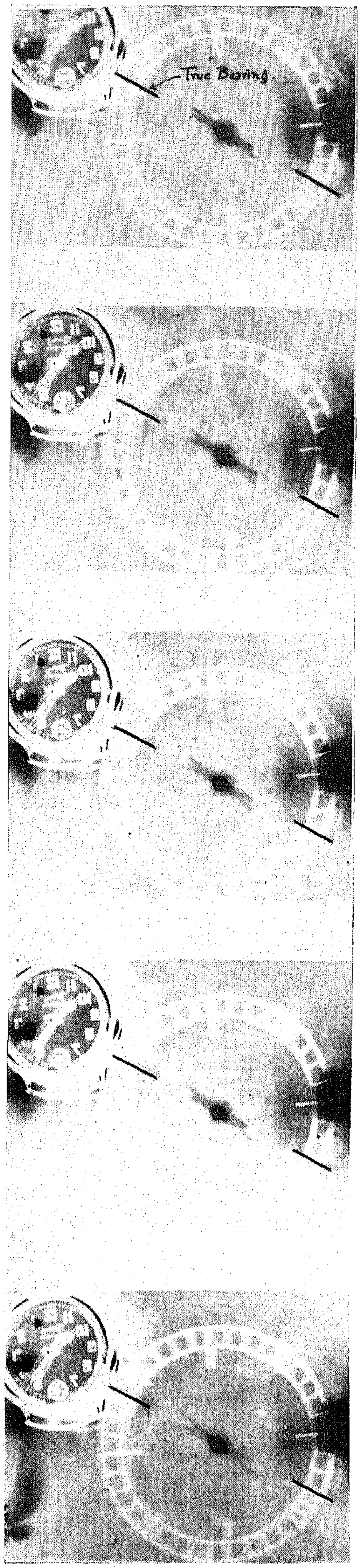

Fig. 7(f).—Dorchester (37·3 m.); 1020 G.M.T.; 23.4.37. Exposure = $\frac{1}{10}$ sec. Interval between photographs = 10 sec. Pulse transmission. Two rays (unidentified) giving different bearings.

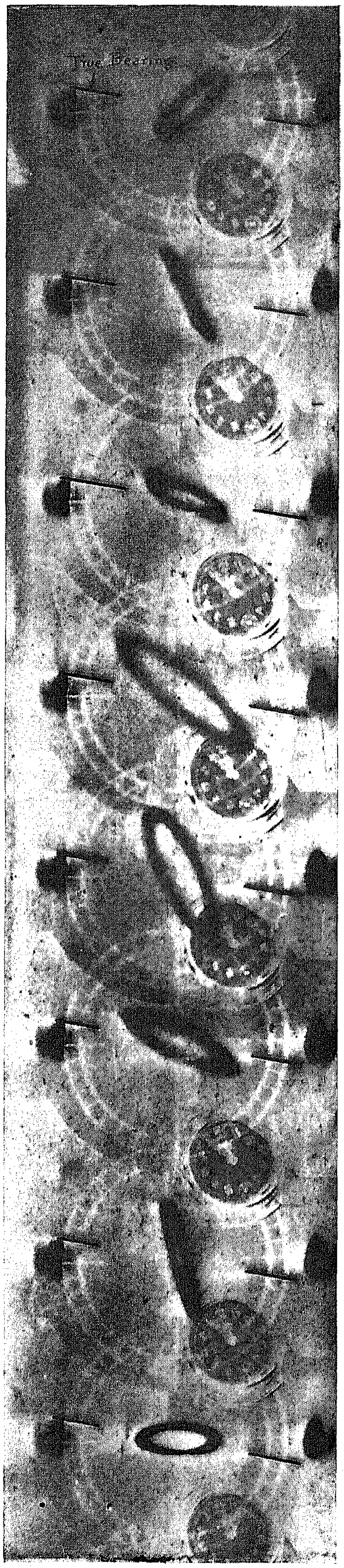

Fig. 7(g).—Zeesen (31·4 m.); 0004 G.M.T.; 18.9.36. Exposure = $\frac{1}{25}$ sec. Interval between photographs = 2 sec. Receiver within skip distance.

Thus, provided note is taken of bearings only when the amplitude of the signal is reasonably large, the instrument itself cannot introduce fluctuations in bearing greater than a maximum of ± 2°, and the probable fluctuations arising from instrumental causes is of the order of ± 1°. In the course of many observations fluctuations outside these limits have frequently been observed. This renders it probable that many of the smaller fluctuations observed do not arise instrumentally, since these larger fluctuations cannot do so.

The rejection of bearings observed when the amplitude is small, further ensures that actual deviations occurring in propagation are not magnified. Such magnification in the apparent deviation of a wave can occur in the case of continuous-wave signals composed of several distinct rays. If each, or any, of the rays is subject to some small amount of lateral deviation, the apparent bearing of the continuous-wave signal will vary as the relative phases of the various rays change, and may suffer apparent deviations several times as great as those of the individual rays. Such spurious effects, however, can again be minimized by observing bearings taken only when the amplitude is large, as mentioned above.

In all results mentioned later in the paper, these conditions have been observed and it is felt that the deviations noted arise predominantly in the process of propagation of the wave.

(4) OBSERVATIONS AND TESTS

The apparatus reached the stage of development recorded in this paper in the summer of 1936, about which time it was felt that it might profitably be taken into use for observations on the direction of arrival of waves.

Bearings were recorded by taking snapshot photographs (usually several per minute) of the fluorescent screen of the cathode-ray tube (see Fig. 7, Plates 1 and 2, facing page 320). By this means it is possible to record bearings practically instantaneously and at regular intervals of time. The exposures used were $\frac{1}{25}$ sec. for continuous-wave signals and about $\frac{1}{5}$ sec. for pulses. Bearings were taken in groups usually extending over about 5 to 15 minutes, the adjustment of the apparatus being checked at the beginning and end of each short period. The results of several different series of observations are recorded below.

(a) Zeesen.

The first series of tests with the apparatus was made by observing the bearings given by the German short-wave broadcasting station at Zeesen. This station is situated 920 km. from the Radio Research Station, and uses a wavelength of 31·4 metres. Observations were made whenever transmissions were available during the months of September and October, 1936, at which period the station was transmitting between the hours 0500 and 1000 G.M.T. and between 2200 and 0300 G.M.T.

It was found that results fell into two distinct classes. At times when the receiver was not within the "skip-distance" from Zeesen, a reasonably definite and steady bearing was obtained [Fig. 7(*a*)]. The bearing was usually subject to continuous and variable oscillation about some mean position quite close to the true bearing, the oscillation being usually within an arc of about 5°; deviations up to about 20°, however, were observed very occasionally when the amplitude was small. Disregarding the signals of smaller amplitude, however, the maximum deviation from the mean which was observed was about 10°. Bearings were usually recorded at the rate of about 12 per minute, those taken when the amplitude was greatest being finally selected from the recorded bearings. In the observations here referred to, all those for which the amplitude fell below 0·7 times the maximum were rejected. The probability that the deviations observed in those results which remain occur in propagation is thus very high.

Table 2 gives the distribution of bearings relative to the mean and is based on about 300 results.

Table 2

ZEESEN (λ = 31·4 METRES, DISTANCE 920 KM.)

Percentage of observations having deviations from the mean not exceeding:—

1°	2°	3°	5°	10°
54	85·5	96	99	100

The observations in each group were averaged and it was found that the group means lay very close to the mean of all the observations. The mean of all the observations had an error of + 2°, which represents the more or less fixed instrumental error. Of the 15 groups of bearings on which the analysis is based, 7 had a mean bearing within 1° of the mean of all the observations, and all were within 2° of that mean. Variations in the state of adjustment of the apparatus could account for variations of the absolute instrumental error of not more than 1°. The results obtained, therefore, show that during these tests there were no occasions on which a steady lateral deviation of more than between 1° and 2° persisted throughout the duration of a group of observations. Table 2 shows, however, that the waves received are subject to continuous changes of bearing of not more than 10° about a mean position which appears to be close to the true bearing. The probable deviation of a single observation is seen to be of the order of 1°.

At times when the receiver was within the skip-distance from Zeesen (from a few hours after sunset until about sunrise) the results obtained were different. Signal strength fell some 30 to 40 decibels. The form of the trace on the oscillograph screen passed rapidly and continuously through all forms of ellipses from a straight line to a circle, the major axis at the same time swinging through all possible orientations [Fig. 7(*g*)]. There was a slight tendency for a predominance of bearings in a direction some 20° from the true bearing, but the bearings recorded were very widely scattered about this direction—about 50 % showing bearings more than 20° different from the means of all the recorded results. The mean bearing indicated agrees very approximately with the bearing to be expected if the scattered signals received at this time came mainly from points along the beam of the transmitter. It is, however, clear that no reliable bearings could be obtained at such times.

(b) Nauen.

To throw further light on the results obtained in the tests mentioned above, a short series of special pulse transmissions from the German station at Nauen (920 km. distant, wavelength 30·6 metres) were arranged with the German Post Office through the kind co-operation of Dr. T. Walmsley of the British Post Office. The transmissions took place in December, 1936, between the hours 1300 and 1400 G.M.T. It was found that with the pulses used ($\frac{1}{8}$ millisec. long) in general two rays were received, corresponding to the first- and second-order reflections from F_2 region. The third-order ray was also received at times, and very occasionally others which may have been reflections from E or F_1 regions [Fig. 7(*d*)].

It was found that all the rays were subject to continuous and variable deviations. In the case of the ray corresponding to the first-order F_2 reflection, the trace on the tube was nearly always linear but occasionally became slightly elliptical, while fading, though not so rapid as in the case of continuous waves, was usually present. This indicates that the ray cannot always be regarded as a single ray but is composite. In view of this, in producing Tables 3 and 4, observations made at small amplitudes have been neglected. The effect on the distribution of errors was found to be slight and much less than in the case of continuous waves. The greater deviations (between 10° and 20°) were, however, eliminated by this process of selection, such deviations being probably occasioned by the interference effects referred to above. The distribution of bearings relative to the mean is given in Table 3, which is based on about 550 results.

Table 3

NAUEN ($\lambda = 30 \cdot 6$ METRES, DISTANCE 920 KM.)

1st-order F_2 reflection

Percentage of observations having deviations from the mean not exceeding:—

1°	2°	3°	5°	10°
49·5	73	85	98	100

The mean of all the observations had an error of − 3°. The observations were, as before, taken in groups at the rate of several per minute over intervals of about 10 minutes. Sixteen such groups of bearings were taken and, of these, 13 groups had a mean error with respect to the mean of all observations of 1° or less, the remaining 3 having relative errors of − 2°, + 2·5°, −3·4°. This indicates that steady lateral deviation of more than 1° persisting for several minutes on end is uncommon, but does occasionally occur and may have an amplitude of 3° to 4°. Individual bearings, however, are subject to a continuously variable amount of lateral deviation up to about 10° maximum, the probable deviation being about 1° as shown in Table 3. The angle of incidence for this ray was measured and found to be about 50°.

Results for the second-order ray were very different from those for the first-order ray. Of 13 groups of bearings obtained on this ray only two groups had mean bearings differing by 1° or less, 5 had mean bearings differing by 2° or less, and 8 had mean bearings differing by 5° or less from the mean of all the observations. The mean bearings of the remaining 5 groups had deviations from the mean of all the observations of − 5·8°, + 6·2°, + 7·2°, − 7·7°, + 8·2°. These groups extended over 5 to 10 minutes and it is seen that, in the case of this ray, steady deviation of one sign might persist for several minutes continuously and might have a maximum value of about 8°. The individual bearings in each group, of course, fluctuate rapidly about the mean bearing for the group. Table 4 shows the distribution of the individual observed bearings relative to the mean of all the observations and has been compiled from 250 results, signals of small amplitude being neglected.

Table 4

NAUEN ($\lambda = 30 \cdot 6$ METRES, DISTANCE 920 KM.)

2nd-order F_2 reflection

Percentage of observations having deviations from the mean not exceeding:—

1°	2°	3°	5°	10°	20°
18	30	42	58	85	99

The probable deviation is seen to be between 3° and 5°. Only a few observations were made on the third-order ray, but these showed that the deviations suffered by this ray were still greater, the probable deviation being between 5° and 10°. The number of results is too small to make the construction of a table of much value. It may be added that the first-order ray was usually much the strongest, being usually at least twice as strong as the second-order ray. Continuous-wave signals from such a station, and in similar circumstances, will therefore consist mainly of the first-order ray, and might be expected to show results similar to those obtained with the first-order ray. The results obtained in the Zeesen tests are in confirmation of this supposition, as a comparison of the results will show.

The results of the Nauen measurements agree on the whole satisfactorily with those obtained by Eckersley* from measurements made with a single-pair spaced-loop arrangement on pulses from the same transmitter.

(c) Prague.

A number of observations have been made on signals from Prague on a wavelength of 44·6 metres at a distance of about 1 040 km. during March and April, 1937, and at all hours of the day between 1000 G.M.T. and 0100 G.M.T. In this case the emissions are morse signals, the carrier wave being modulated. The figure on the cathode-ray-tube screen was frequently extremely complex, several thin ellipses being seen at once in addition to other fainter ellipses apparently corresponding to key-clicks [Fig. 7(*c*)]. The bearings indicated by the different ellipses were generally quite different. On some occasions only one ellipse was visible. When several ellipses were present, the bearing of the one of greatest amplitude only has been taken into account, all signals whose amplitude fell to below 0·5 times the maximum being rejected. The results obtained are similar to those obtained on Zeesen, although the deviations noted are somewhat greater.

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* See Bibliography, (13).

The mean of all the observations had an error of — 4°. Of 27 groups of bearings recorded, 15 had mean bearings within 1° and 23 had mean bearings within 2° of the mean of all. The remaining 4 groups had deviations from the mean of 2·5°. The nature of the results was practically the same at all times when transmissions were observed, and at no time did the receiver pass into the skip region from Prague. The distribution of bearings for all the individual observations noted is shown in Table 5, which is based on 700 observations. There were occasions when the distribution was better than that shown in Table 5, such times usually occurring during the middle of the day. There are insufficient data, however, to correlate any particular type of results with a particular time of day, except to note that all four occasions on which the mean relative error of a group exceeded 2° occurred in the late evening.

Table 5

PRAGUE (λ = 44·6 METRES, DISTANCE 1 040 KM.)

Percentage of observations having errors relative to the mean not exceeding:—

1°	2°	3°	5°	10°	20°
39	58	72	85	98	99·7

The results appear to show that the waves received were quite frequently subject to steady deviations of about 1° to 2°, since 12 out of 27 groups had deviations of more than 1° from the mean of all the results. Such steady deviation, however, was not found to exceed 2·5°. The probable deviation of a single observation is between 1° and 2°.

(d) Various American Fixed Stations.

Observations were made during October and November, 1936, on stations situated at Sayville (wavelength 36 metres), Chatham, Mass. (36·5 metres), Millis, Mass. (31·4 metres), Schenectady (31·5 metres), and New Brunswick, N.J. (31·7 metres), all some 5 000 km. from the Radio Research Station.

Observations were made when transmissions were available and could be picked up at sufficient strength, namely between the hours of 2000 G.M.T. and 0400 G.M.T. It was found that deviations were usually somewhat larger than those observed on the stations already mentioned, and the tendency for a steady deviation to persist for several minutes was somewhat greater [Fig. 7(*b*).] Thus of the 9 groups of observations made, only 5 had mean bearings differing by 2° or less from the mean of all, and the mean of one group had a deviation of 4·6° relative to the mean of all. It has been assumed that, since all the above stations lay in approximately the same direction and used wavelengths quite close to each other, the fixed error would be approximately the same for all of them. This fixed error appeared to be about — 3°. The distribution of bearings for individual observations is given in Table 6, which is based on about 250 observations. About 500 of the observations actually made were rejected as having too small an amplitude. The nature of the results was inclined to be very variable, bearings being occasionally much steadier than Table 6 implies, and at times even less steady. The probable deviation of a single observation appears to be between 2° and 3°.

Table 6

VARIOUS U.S.A. TRANSMITTING STATIONS (WAVELENGTH 31·4 TO 36 METRES, DISTANCE 5 000 KM. APPROX.)

Percentage of observations having errors relative to the mean not exceeding:—

1°	2°	3°	5°	10°	20°
24	39	56	77	95	100

These results again may be compared with those obtained by Eckersley* from a few observations made on Montreal (λ = 32·15 m.) at Chelmsford with his single-pair spaced-loop system, when deviations of the same order were recorded.

(e) Dorchester.

Observations have been made on pulse transmissions from Dorchester on a wavelength of 37·3 metres during the months of April, May, and June, 1937. This station is about 160 km. from the receiver. The signals which are received are, therefore, incident at quite a small angle to the vertical. For the first-order reflection from the F_2 region, the angle of incidence of the ray was probably between 12° and 18° during the period of the tests. At such angles of incidence the instrumental accuracy is, of course, not of such a high order as for rays at much greater angles of incidence, since the effective spacing between the aerials is less, and small misadjustments of the aerials may produce quite appreciable errors. This affects most particularly the absolute accuracy of the direction-finder. The minor fluctuations in aerial balance which occur more or less continuously during observations also produce correspondingly larger fluctuations in the apparent bearing. Any residual polarization errors also become increasingly great as the angle of incidence decreases. For equal horizontally and vertically polarized components in the wave, however, the polarization error for an angle of incidence of about 12° is not greater than about 1°. It will increase, of course, as the state of polarization of the wave changes in such a manner as to reduce the wanted pick-up of the aerials, but provided bearings are only noted when the amplitude of the signals is near the maximum the error cannot exceed, say, 2° at most. Tests have shown that for signals from this station the average strength of the horizontally and vertically polarized components are equal. At the most, therefore, instrumental errors cannot give rise to variable errors in the bearing of more than 4° or 5°, and probably the instrumental variations do not cause any deviations greater than about 2° or 3°.

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* See Bibliography, (13).

The results obtained in the course of tests on Dorchester are, in the main, strikingly different from those obtained on the more distant stations and are most conveniently illustrated by graphs. This has been done in Fig. 6. For any particular ray the bearing fluctuates about some mean position, but this mean position is frequently many degrees from the true bearing. It has been observed to be as much as 50° to 60° in error, and moreover is frequently observed to drift steadily in one direction or the other. Thus a continuous change in the mean bearing of as much as 25° in 10 minutes has been observed (Fig. 6, 30.4.37, 0940 G.M.T.). Occasionally, too, the drift may change direction after several minutes, so that the mean bearing returns to its original value (Fig. 6, 30.4.37, 0940 G.M.T.). The mean bearing has also been observed to oscillate slowly about the true value (Fig. 6, 23.4.37, 1020 G.M.T.). The fluctuations of the bearing about its instantaneous mean position are usually of small amplitude, i.e. ± 3° or 4° at most. Moreover, when more than one ray is present the two rays frequently show very different bearings, as much as 10° to 20° difference in bearing being observed for the first- and second-order reflections from F_2 (Fig. 6, 21.5.37, 1030 G.M.T.). At times also the first-order reflection has been observed to be split into its two components by magneto-ionic action. When this occurs the two components have been observed sometimes to arrive in exactly the same direction and to continue to do so for several minutes together, although that direction of arrival may not correspond to the true bearing (Fig. 6, 4.6.37, 0930 G.M.T.—first 7 minutes), while at other times they may have bearings differing by up to 30° (Fig. 6, 7.5.37, 1050 G.M.T., and 4.6.37, 0930 G.M.T.—last 10 minutes). It should be noted that the effect of any substantial polarization error in the direction-finder would be to produce a difference in apparent bearing for these two components of the first-order F_2 reflection of more or less constant amplitude and of constant sign and such that the mean bearing of the two was approximately true. This is deduced from the known fact that the two components will be nearly circularly polarized, having opposite senses of rotation. The results show, however, that the difference in bearing for the two components can be anything from zero to 30° and is not of constant sign, sometimes the one and sometimes the other being most northerly. The mean bearing may not be even approximately true.

On a few occasions, when a single ray only has been received, results have been very erratic, the bearing changing by 20° or 30° from one instant to the next and not showing any tendency to indicate a mean direction (Fig. 6, 4.6.37, 1100 G.M.T.). This might be expected to occur at times when the ray consisted of several components (such as the two magneto-ionic components, not actually separated in time) each of which was subject to a large amount of deviation. As the phase relationship

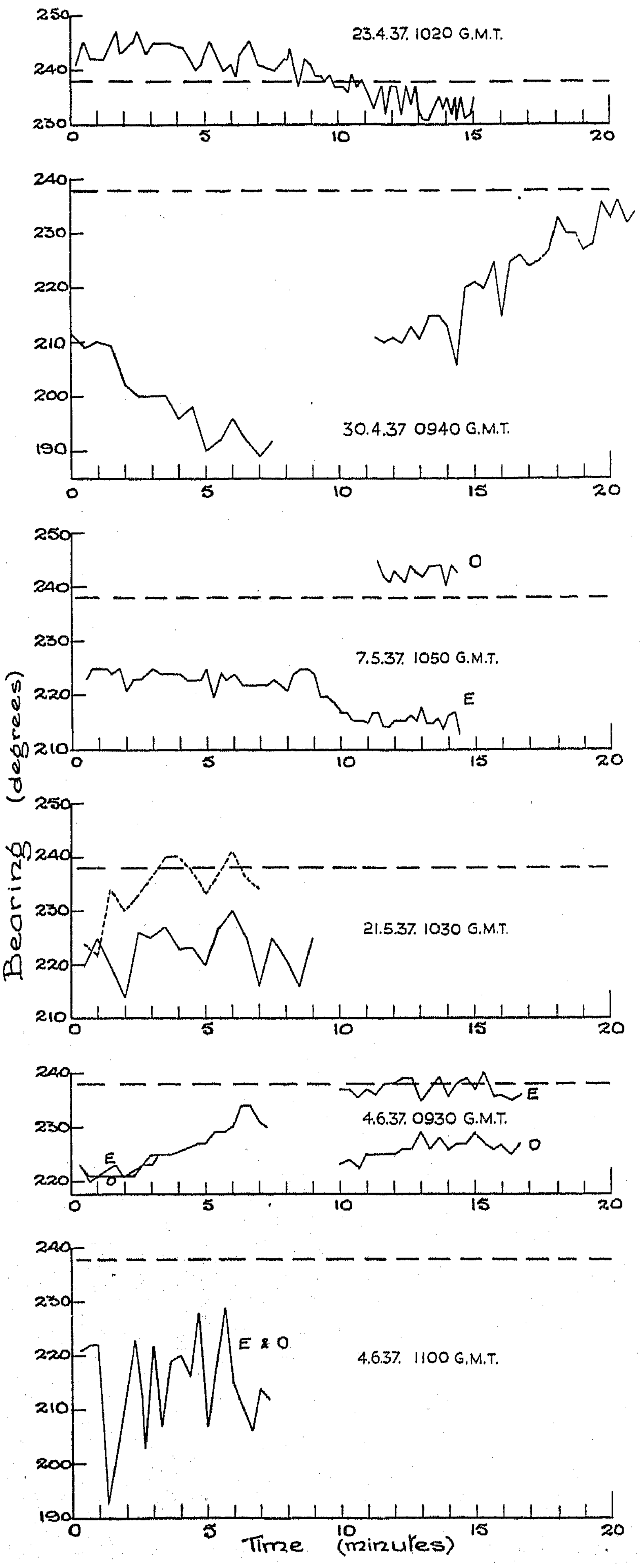

Fig. 6.—Observations on pulse transmissions from Dorchester (λ = 37·3 m.).

— — — True bearing.
——— 1st-order F_2 reflection.
- - - - - - 2nd-order F_2 reflection.

O = Ordinary component.
E = Extraordinary component.

of the components changed, the apparent bearing would fluctuate to a very great extent. At such times also, the trace, which is usually practically linear, may become very much opened out into an ellipse.

These results can only be interpreted as implying that considerable lateral deviation of waves which have travelled by way of the ionosphere can exist. Instrumental errors might possibly account for the small fluctuations of the bearing from instant to instant in the case of waves incident at such small angles to the vertical as those received from Dorchester, but cannot account for the large deviations of up to 50° which have been observed. Such deviations imply that the region of the ionosphere effective in reflecting the waves received at the direction-finder may be displaced laterally by as much as 50 to 100 km. from the great-circle path joining the transmitter and receiver.

(5) CONCLUSIONS

It has been shown that the spaced-loop direction-finder is to a very large extent free from instrumental errors of a variable nature, and that, provided care is taken in rejecting bearings obtained when the amplitude is small, the variations in apparent bearing which can arise from instrumental causes have a maximum of about 2° and are probably usually of the order of 1°. The results of various tests made with the apparatus show that deviations in apparent bearing much in excess of these figures are found, deviations up to 10° or 20° being noted on distant stations and up to 50° on a station at close range. It seems practically certain that these deviations arise in propagation and that, in fact, the wave does not travel along the great-circle path. It appears that the equivalent point of reflection at the ionosphere may be as much as 50–100 km. out of the great-circle path. Usually the deviations suffered by waves change rapidly from instant to instant in such a way that the mean bearing is very close to the true one, but occasions are not rare on which the mean direction of arrival of the waves remains deviated in one direction for several minutes on end.

Waves which have had several reflections at the ionosphere appear to be liable to greater deviation, as might be expected. This is borne out by direct observations on waves known to have suffered two reflections, and by observations on certain American stations for which all the waves must have suffered multiple reflections at the ionosphere.

There is some indication that so-called "individual" rays are perhaps something in the nature of cones of rays. This deduction is based on the fact that, when receiving pulses, the traces for the "individual" rays are quite often slightly elliptical. This can only occur if more than one ray is present in each so-called "ray" and the constituent waves do not arrive along precisely the same path. The direction indicated will be some mean direction, usually within the cone of rays except when the rays interfere so as to produce a signal of very small amplitude, when the apparent direction may lie quite outside the cone. The variations in apparent bearing from instant to instant may therefore be due, not so much to instantaneous changes in the direction of arrival of the rays, but to phase changes amongst the components of the ray, giving rise to changes in the apparent bearing and degree of ellipticity of the trace. It is clear, too, that if this interpretation of the results is true, the whole cone of radiation constituting a ray may, and apparently does, suffer deviation from the great-circle path.

The experience gained with the apparatus described in this paper has indicated certain directions in which the direction-finder might be further improved so as to attain a higher degree of precision and certainty in the measurement of small amounts of lateral deviation. An improved model of the direction-finder is therefore at present being designed, with which it is believed that the problem of lateral deviation of the ionospheric ray may be adequately attacked.

(6) ACKNOWLEDGMENTS

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COEXISTENT THERMAL AND THERMIONIC FLUCTUATIONS IN COMPLEX NETWORKS[*]

By F. C. WILLIAMS, M.Sc., D.Phil., Graduate.

(*Paper first received 2nd December*, 1937, *and in final form 16th February*, 1938.)

SUMMARY

This paper is concerned with the fluctuations generated in networks with non-uniform temperature distribution. Experiments are described which verify a formula relating to thermal fluctuations deduced in an earlier paper. The simplified representation adopted in that paper is fully substantiated: the fluctuations are shown to be independent of the temperature of reactive circuit elements.

Coexistent thermal and thermionic fluctuations are then discussed. They are assumed to be mutually independent, and experiments are described which verify this hypothesis. A representation of thermionic fluctuations is proposed analogous to that adopted for thermal fluctuations, and a general formula applicable to certain classes of networks containing thermionic valves is deduced.

This formula is applied in a concluding section to a wide-band photocell amplifier. Various practical recommendations are made which yield improved signal/noise ratios.

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(1) INTRODUCTION

In a recent paper[†] the author deduced a general formula for the thermal fluctuations[‡] appearing between any two points in a network of linear conductors. The analysis was based on the hypothesis that the fluctuations can be regarded as being produced by elementary fluctuation generators supposed acting in series with each element of resistance. Thus in series with the typical element of resistance r_x there was supposed to exist a generator delivering a mean-square voltage within any frequency band *df* of value $\overline{v_x^2} = 4r_xkT_xdf$.

It was then shown that the mean-square voltage, within the frequency band *df*, appearing between any two points B and B′ of the network was

$$\overline{v^2} = Z_{BB'}^2 4kdf \sum \frac{r_xT_x}{Z_{Bx}^2} \quad . \quad . \quad . \quad . \quad (1)$$

where

Z_{Bx} = modulus of the transfer impedance from the hypothetical generator in series with the typical resistance, r_x, to a short-circuit joining the points B, B′. That is $1/Z_{Bx}$ = current in the link joining BB′ per unit e.m.f. in series with r_x.
$Z_{BB'}$ = modulus of the network impedance from B to B′.
k = Boltzmann's constant.
T_x = absolute temperature of r_x.

The impedance values are relevant to the mid-frequency of the small band *df*.

With uniform temperature distribution this was shown to reduce to

$$\overline{v^2} = 4kTrdf \quad . \quad . \quad . \quad . \quad . \quad (2)$$

where r is the real part of the impedance from B to B′, a result in accordance with Nyquist's earlier analysis.[†]

Equation (2) was verified experimentally in the earlier paper: one purpose of the present paper is to verify the general equation (1).

Llewellyn[‡] has used the equation

$$\overline{v^2} = \left(\frac{R_1R_2}{R_1 + R_2}\right)^2 4kdf\left(\frac{T_1}{R_1} + \frac{T_2}{R_2}\right) \quad . \quad . \quad (3)$$

relevant to a resistance R_1 at temperature T_1 in parallel with a resistance R_2 at temperature T_2. This is a particular example of the general equation (1). Moullin and Ellis[§] have described an experimental check of this equation: their results were not in perfect agreement with it, but they ascribed the discrepancy to imperfections of their experimental method and did not then consider that it merited closer investigation.

The present paper also extends equation (1) to cover certain classes of networks generating thermionic as well as thermal fluctuations.

This extension has been made in order to facilitate the prediction of the total noise output of amplifiers: the well-known simple formulae for shot and thermal effects permit such prediction only when one or other of the effects is dominant, for the mechanism of their addition in complex networks has not been studied. The non-uniform temperature distribution is included partly for the sake of completeness, but also because variation of the temperature of a resistance modifies its thermal fluctuation without modifying the circuit parameters, thus facilitating the experimental determination of the contribution of each circuit element to the total fluctuation.

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[*] Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 76.

[†] See Bibliography, (1). (B and B′ replace A and A′ in the paper referred to.)

[‡] The term "fluctuation" is here used to refer to the spontaneous oscillations of potential which appear in electrical circuits due to "Brownian motion," "shot effect," etc.

[†] See Bibliography, (2).

[‡] *Ibid.*, (4).

[§] *Ibid.*, (3).

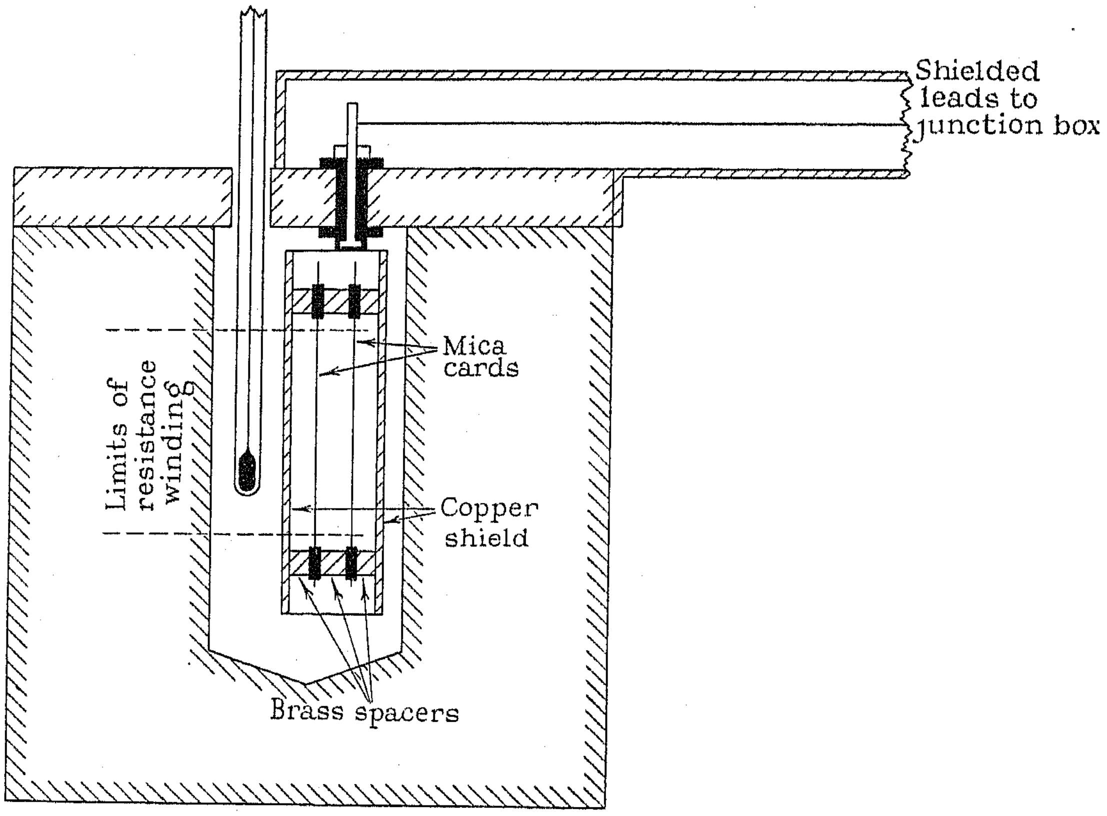

Fig. 1.—Sketch of box containing heated resistance. Mica washers shown black.

It may appear that equation (1) does not strictly relate to circuits with "non-uniform temperature distribution," but only to circuits having finite elements at different temperatures. However, the elements of resistance such as r_x can be taken very small and the condition of continuous temperature variation approached as closely as desired.

An example of the use of the formulae is given in a concluding section, where they are used to determine the noise/signal ratio in a wide-band photocell amplifier.

(2) EXPERIMENTAL APPARATUS

An 8-stage amplifier supplying a thermojunction was used; its design was straightforward and does not merit detailed description. The response characteristic was adjustable by the insertion of suitable filter circuits. The magnification could be varied by adjustment of screen grid potentials: at its maximum, full-scale deflection (50 cm.) corresponded to an input voltage of about $0 \cdot 1\ \mu$V (r.m.s.). Fluctuations generated in the amplifier itself were equivalent to the thermal-agitation voltage generated in a resistance of $2 \cdot 3$ kΩ connected to the input terminals.

A junction box of $\frac{1}{8}$ in. brass sheet with a flush-fitting lid contained that portion of the apparatus which was to be kept at room temperature. The output from a calibrated attenuator was fed to this box, as also were the input leads from the amplifier, and the shielded leads from the box containing the circuit elements to be heated.

The box containing the heated elements is shown in Fig. 1. It consisted of a 6-in. cube of cast iron having a 2 in. dia. hole drilled as shown; the cast-iron lid was $\frac{1}{2}$ in. thick and the leads passed out through mica bushes. This massive construction was adopted in order to obtain a substantially uniform temperature inside the box, so that the measured temperature should be that of the resistance. The box was heated externally with bunsen burners, and was allowed to attain a steady temperature before each set of observations was taken.

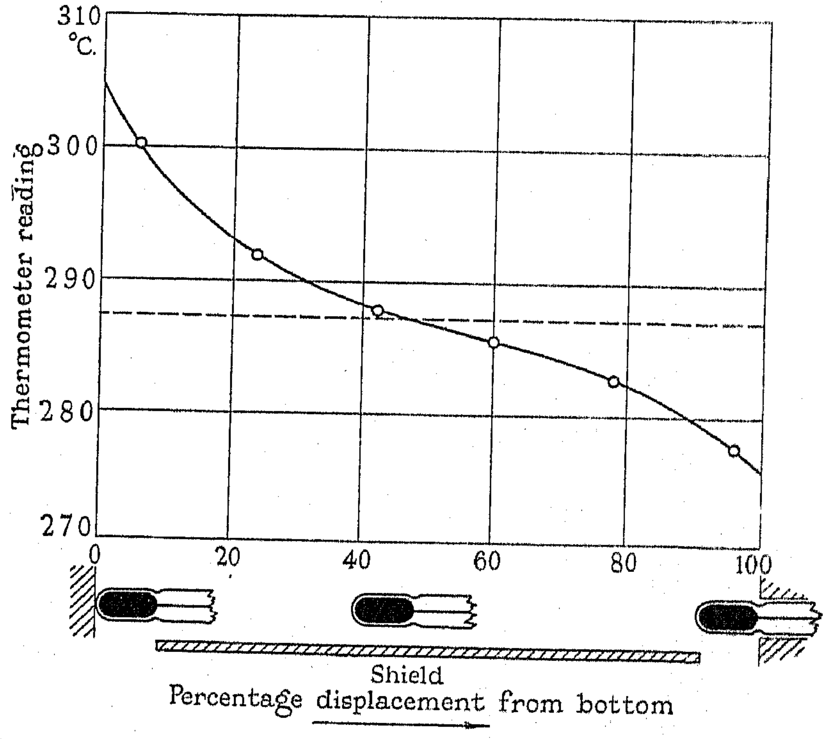

Fig. 2.—Vertical temperature distribution inside the heated container shown in Fig. 1.

A test of temperature distribution is shown in Fig. 2, where recorded temperature is plotted as a function of the vertical immersion of the thermometer: it can be seen that the whole range of temperature is only 29 deg. C.; that is, little more than 10 % of the excess of the mean temperature above the room temperature. It is presumed that this figure, 10 % of the excess, is sensibly independent of the excess, since the non-uniformity is due to uneven heating of the box, which was always heated from below. Thus with a maximum temperature of 500° C. the range of temperature probably had a maximum value of about 7 % of the absolute temperature (50° K. in 773° K.). The resistances to be heated were wound in two sections on thin mica cards. The wire was 0·001 in. dia. nichrome. The two cards were enclosed in a copper shield and spaced from each other and from the shield by transverse lengths of $\frac{1}{4}$ in. square brass rod. The spacers, together with the shield, gave a totally enclosed resistance (see Fig. 1). The temperature inside the shield must therefore have been very close to the average temperature over its outside surface. The positions of the shield and thermometer are shown in Fig. 1; the correct immersion for the thermometer was deduced from Fig. 2. On the basis of Fig. 2 and the foregoing discussion it is thought that errors due to uneven temperature distribution cannot have exceeded 1 % in absolute measure.

The thermometer was of quartz and was calibrated with boiling water, naphthalene, benzophenone, and sulphur. The calibration points were all within 4 % of the marked temperature.

(3) EXPERIMENTAL PROCEDURE

The input impedance of the amplifier was of the order of 0·25 MΩ, and was believed to be largely capacitative. The effect of this on the fluctuations generated by the test apparatus was eliminated by the technique described below. In Fig. 3, $\mathbf{Z} = R + jX$ represents the complex test apparatus, while jX' represents the reactive input impedance of the amplifier and its associated leads.

The sinusoidal calibrating voltage V of frequency f was introduced at C.

It follows from equation (1) that the mean-square fluctuation voltage between B and B′ in the narrow frequency range df situated about f is

$$\overline{v^2} = \left|\frac{\mathbf{Z}jX'}{\mathbf{Z}+jX'}\right|^2 4kdf \sum \frac{r_xT_x}{Z_{Bx}^2}$$

The mean-square voltage from B to B′ due to V is

$$\overline{v_c^2} = \left|\frac{jX'}{\mathbf{Z}+jX'}\right|^2 V^2$$

The ratio of these expressions is

$$\frac{\overline{v^2}}{\overline{v_c^2}} = \frac{\left|\dfrac{\mathbf{Z}jX'}{\mathbf{Z}+jX'}\right|^2}{\left|\dfrac{jX'}{\mathbf{Z}+jX'}\right|^2} \cdot \frac{2kdf}{V^2} \sum \frac{r_xT_x}{Z_{Bx}^2}$$

This ratio is independent of X': thus the reactive component of input impedance does not affect the measurement. A comparison of the deflection due to V and that due to fluctuations generated in $\mathbf{Z}$ and jX' in parallel thus measures the fluctuations which would appear across $\mathbf{Z}$ if it were isolated. This result holds only if the amplifier input impedance is purely reactive; it can be shown that in the general case the "series" method of comparison is preferable to that in which first $\mathbf{Z}$ and then V are connected to the amplifier, the ratio of the two deflections being taken as a measure of the fluctuation generated in $\mathbf{Z}$ if it were isolated.

The amplifier was not calibrated in absolute measure. The fluctuations generated in the test apparatus were compared with those generated in a metallic resistance of known value. The observed fluctuations were expressed in terms of the resistance which generated a fluctuation equal to that generated in the test apparatus.

The following deflections were noted:—

θ_1 . . . with the amplifier input short-circuited. This was due to amplifier noise.
θ_2 . . . with the test apparatus connected to the amplifier.
θ_3 . . . as θ_2 but with a sinusoidal calibrating voltage V in series with the test apparatus.
θ_1 . . . repeated.
θ_4 . . . with a metallic resistance R' connected to the amplifier.
θ_5 . . . as θ_4 but with V in series with R'.

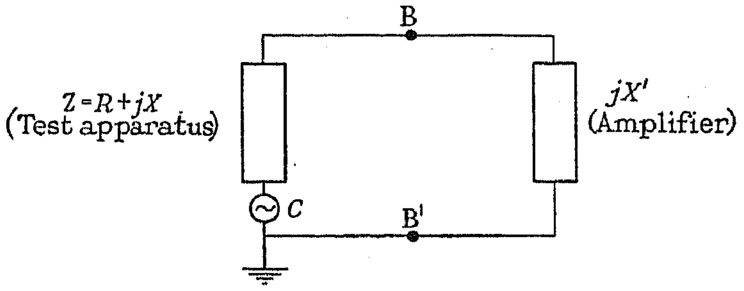

Fig. 3

The deflections were not steady, but showed a random variation of about 5%. Accordingly, 6 sets of readings $\theta_1 \ldots \theta_5$ were taken for each experimental point, values of $(\theta_2 - \theta_1)/(\theta_3 - \theta_2)$ and $(\theta_5 - \theta_4)/(\theta_4 - \theta_1)$ being derived from each set.

Any individual value of either ratio which differed more than 5 % from the mean was rejected as spurious. Such rejections averaged about 1 in 10; the bulk of the remainder were within 2 % of the mean.

By taking measurements in this order, the effect of the inevitable slow decrease of amplification with time was eliminated, for each set of measurements, θ_1 to θ_3 and θ_1 to θ_5, was taken in less than 2 minutes.

It follows from the above discussion that the measured value of the fluctuation voltage generated across the test apparatus, when isolated, expressed in equivalent kilohms at the temperature of R', is

$$R_e = \frac{\theta_2 - \theta_1}{\theta_3 - \theta_2}\,\frac{\theta_5 - \theta_4}{\theta_4 - \theta_1}\,\mathrm{R}'\ \mathrm{k}\Omega\ (R' \text{ in kilohms})$$

The final value of the observed fluctuations was deduced from this equation, using the mean values of $(\theta_2 - \theta_1)/(\theta_3 - \theta_2)$ and $(\theta_5 - \theta_4)/(\theta_4 - \theta_1)$.

As a test of the method and the apparatus, several new resistances were checked against one which had been shown* to yield a pure thermal fluctuation. Those resistances which were to be heated in subsequent experiments were tested in the special container; the remainder were placed in the junction box for test.

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* See Bibliography, (1) (Section 4).

Fig. 4 shows the results of the test. Here the equivalent resistance in kilohms determined by fluctuation measurements is plotted as a function of the d.c. resistance in kilohms. A straight line having a slope of 45° is drawn on the figure. All points lie close to the line and show a mean divergence of 0·2 % only.

It is to be noted that any appreciable grid current in the first amplifier stage would yield a parabolic curve. The experimental results show no such tendency over the range covered, and it is presumed that grid current was neglibible, i.e. less than, say, 0·05 μA.

(4) RESULTS WITH METALLIC RESISTANCES IN PARALLEL

In this experiment the test impedance **Z** consisted of two sensibly pure resistances. The temperature of one resistance (R_2) could be adjusted to any value between 20° C. and 470° C.: the other (R_1) remained at room temperature.

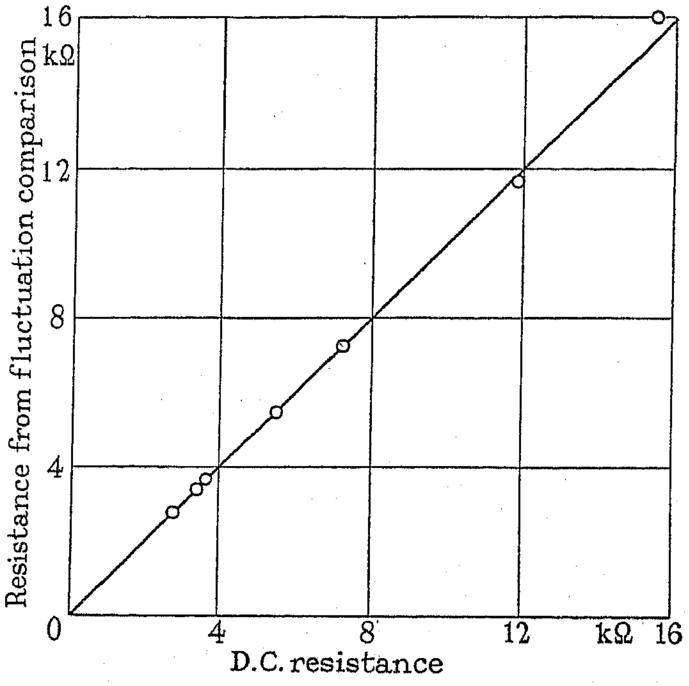

Fig. 4.—Comparison of resistance values obtained from fluctuation measurements with the d.c. values.

It follows from (3) that the relevant fluctuation in the range df is

$$\overline{v^2} = 4R_1kT_1df\left(\frac{R_2}{R_1 + R_2}\right)^2\left(1 + \frac{T_2}{T_1}\cdot\frac{R_1}{R_2}\right)^{*}$$

Expressing $\overline{v^2}$ in terms of the equivalent metallic resistance R_e at the temperature T_1 required to yield an equal fluctuation, we have

$$R_e = R_1\left(\frac{R_2}{R_1 + R_2}\right)^2\left(1 + \frac{T_2}{T_1}\cdot\frac{R_1}{R_2}\right) \quad . \quad . \quad (4)$$

This expression is linear in T_2 if T_1 and R_2 remain constant. The value of R_2 was found to increase 2 % at 470° C.: in presenting the results its value has been taken as independent of temperature and equal to its measured resistance at 220° C. The proximity of the

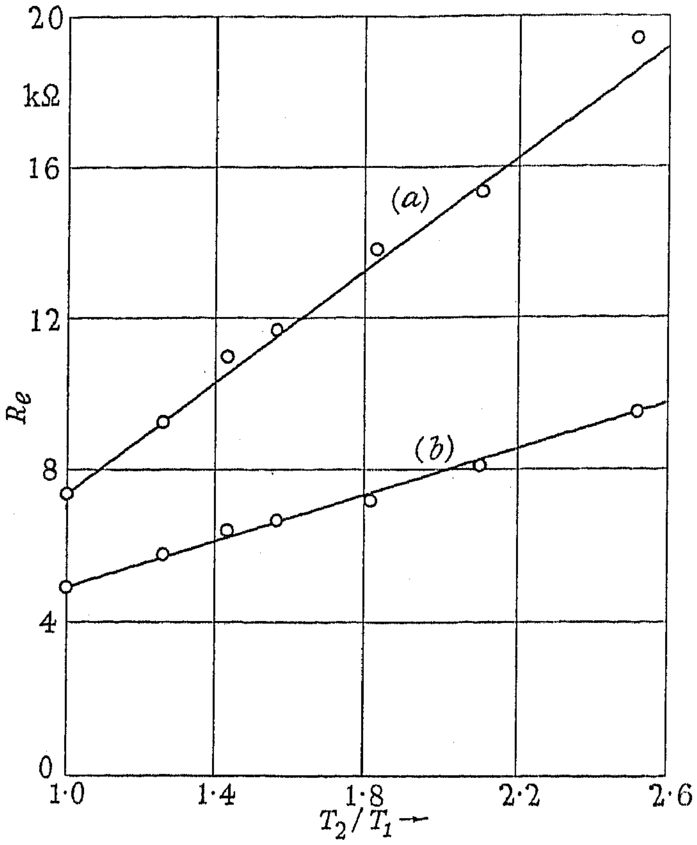

Fig. 5.—Fluctuations across two resistances in parallel as a function of the ratio of the temperatures of the resistances. Amplifier response limited to a 10 % band about a frequency of 8 kc.

Full lines show predicted values.
Plotted points show experimental results.

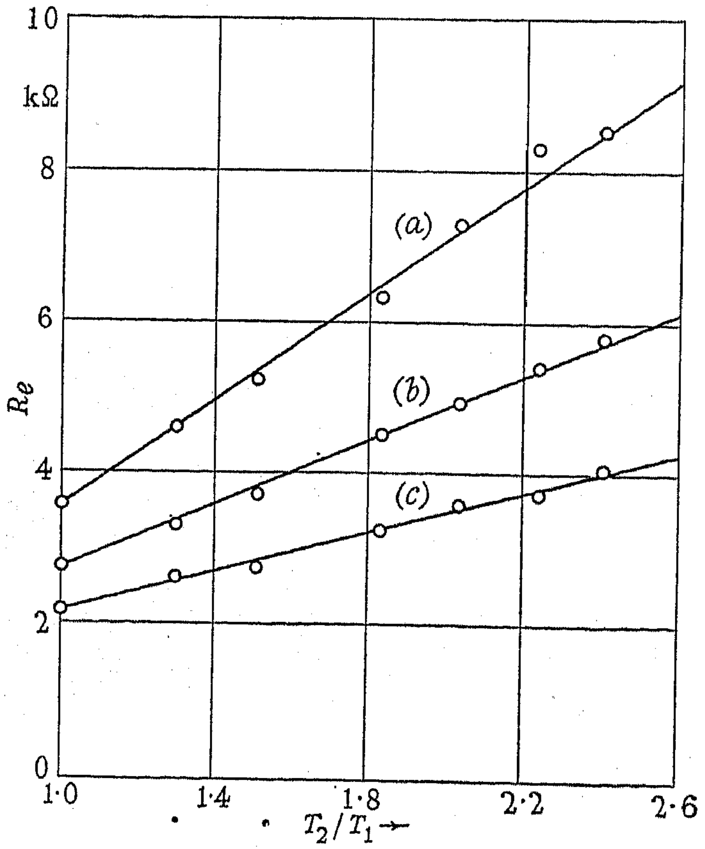

Fig. 6.—As Fig. 5, but with different resistances and with the amplifier response limited to a 5 % band about a frequency of 5 kc.

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* T_2 and T_1 are the absolute temperatures of R_2 and R_1, respectively.

heated apparatus caused T_1 to vary over a range of 5 %: this variation was taken into account in the ratio T_2/T_1.

The results shown in Fig. 5 were obtained with the amplifier passing a 10 % band* centred on a frequency of 8 kc. Line (*a*) on the figure is drawn according to equation (4), with $R_1 = \infty$ and $R_2 = 7{\cdot}41\ \text{k}\Omega$: line (*b*) refers to $R_1 = 13{\cdot}55\ \text{k}\Omega$ and $R_2 = 7{\cdot}41\ \text{k}\Omega$. The observation points lie close to these lines: thus for line (*a*) the maximum discrepancy is + 4 % and the mean is + 0·4 %: for line (*b*) the figures are ± 3 % and + 0·02 % respectively.

Fig. 6 records the results of a similar test, but the amplifier now responded to a 5 % band centred on a frequency of 5 kc. Line (*a*) relates to $R_1 = \infty$ and $R_2 = 3{\cdot}54\ \text{k}\Omega$, line (*b*) relates to $R_1 = 11{\cdot}9\ \text{k}\Omega$ and $R_2 = 3{\cdot}54\ \text{k}\Omega$, whilst for curve (*c*) $R_1 = 5{\cdot}485\ \text{k}\Omega$ and $R_2 = 3{\cdot}54\ \text{k}\Omega$. The maximum and mean discrepancies for lines (*a*), (*b*), and (*c*) are + 4 % and − 0·2 %, − 3·5 % and − 0·6 %, and + 3 % and − 0·3 % respectively.

These experiments appear to establish equation (3) beyond doubt, and verify the view expressed by Moullin and Ellis that the discrepancy they found was due to experimental error.

(5) THERMAL FLUCTUATIONS IN A RESONANT CIRCUIT WITH NON-UNIFORM TEMPERATURE

In the earlier paper† a resonant circuit such as is shown in Fig. 7 was tested under conditions of uniform temperature distribution and shown to yield results supporting equation (1). With the two resistances at different temperatures, equation (1) gives

$$\overline{v^2} = Z_{BB'}^2 4k\,df\left[\frac{R_1T_1}{R_1^2 + L^2\omega^2} + \frac{R_2T_2}{R_2 + 1/(C^2\omega^2)}\right]$$

Hence, if R_1 and R_2 are small compared with $L\omega$ and $1/(C\omega)$, we have, by the process previously adopted,

$$R_e = Z_{BB'}^2 R_1\left[\frac{1}{L^2\omega^2} + \frac{R_2T_2}{R_1T_1}\cdot C^2\omega^2\right]$$

$$= Z_{BB'}^2\frac{R_1}{L^2\omega^2}\left[1 + \frac{R_2T_2}{R_1T_1}\cdot\left(\frac{\omega}{\omega_0}\right)^4\right] \quad . \quad . \quad (5)$$

where $\omega_0 = 1/\sqrt{(LC)}$.

The curves of Fig. 8 show experimental results obtained with a circuit with $C = 0{\cdot}001\ \mu\text{F}$ (nominal) and $L = 183$ mH. R_2 was heated; it had a value of 1 975 Ω, whilst R_1 was 1 860 Ω. Experiment at the resonant frequency showed that residual losses in the coil were equivalent to a series resistance of 390 Ω: this was regarded as independent of frequency. The total value of R_1 was thus 2·250 kΩ. For curve (*a*) ω/ω_0 was unity, for curve (*b*) ω/ω_0 was 1·15, and for curve (*c*) ω/ω_0 was 0·895.

All points lie close to straight lines in accordance with (5).

Values of $Z_{BB'}R_1/(L^2\omega^2)$ were deduced from the observations with $T_2/T_1 = 1$, a condition in which the expression has already been checked.* The dotted lines were then calculated from equation (5). They lie close to the experimental lines, the maximum divergence being 5 %, but they are in all cases steeper by about 10 %.

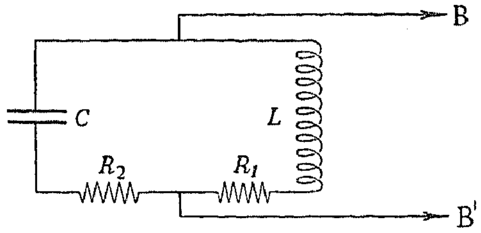

Fig. 7

Fig. 9 records the results of a similar test, but with the resistance (now termed R_2) in series with L heated instead of that in series with C (now termed R_1).† The coil remained at room temperature and thus its inherent resistance R_2' (= 390 Ω) remained at T_1. With these conditions it follows that

$$R_e = Z_{BB'}^2 R_1 C^2\omega^2\left[1 + \frac{R_2'}{R_1}\left(\frac{\omega_0}{\omega}\right)^4 + \frac{R_2T_2}{R_1T_1}\left(\frac{\omega_0}{\omega}\right)^4\right]df$$

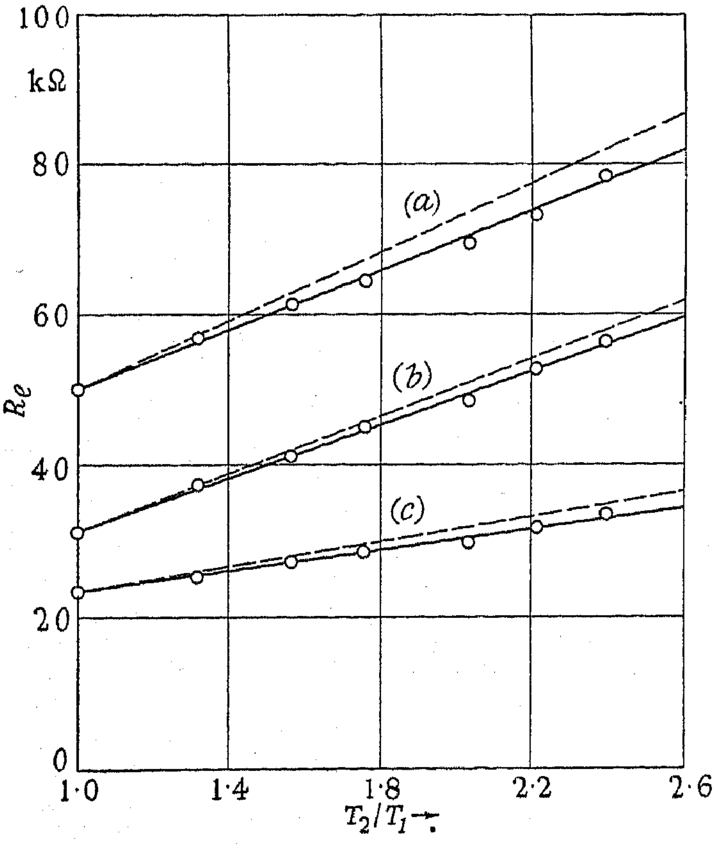

Fig. 8.—Fluctuations in a resonant circuit as a function of the temperatures of the resistance elements.

Curve (*a*).—$\omega/\omega_0 = 1$
Curve (*b*).—$\omega/\omega_0 = 1{\cdot}15$
Curve (*c*).—$\omega/\omega_0 = 0{\cdot}895$
$\omega_0 = 2\pi \times 10\,450$
The amplifier band width was about 3 %.
The dotted lines show the expectation.

The lines in Fig. 9 have been drawn according to this expression, absolute values again being based on the observations with $T_2/T_1 = 1$. The experimental points follow the lines very closely.

Finally the condenser alone was heated, the remainder of the circuit (i.e. L_1, R_1, R_2, and R_2') remaining at room temperature. The amplifier response curve was centred on the resonant frequency of the circuit and had a much greater band-width. Accordingly, the amplifier was sensitive to the whole fluctuation output of the test circuit. The results, recorded in Fig. 10, show conclusively that the fluctuation output is independent of the temperature of C, and hence that the resistance components alone are the seat and origin of the mechanism producing the thermal agitation fluctuation.

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* The "band" is here defined as the percentage frequency interval enclosed between ordinates of the response curve having values $1/\sqrt{2}$ of the maximum ordinate: in comparisons of this kind the exact shape of the response curve is immaterial.

† See Bibliography, (1).

* See Bibliography, (1) (Section 5).

† Thus the definitions of R_1 and R_2 implicit in Fig. 7 are interchanged: this has been done so that T_2 remains the adjustable temperature.

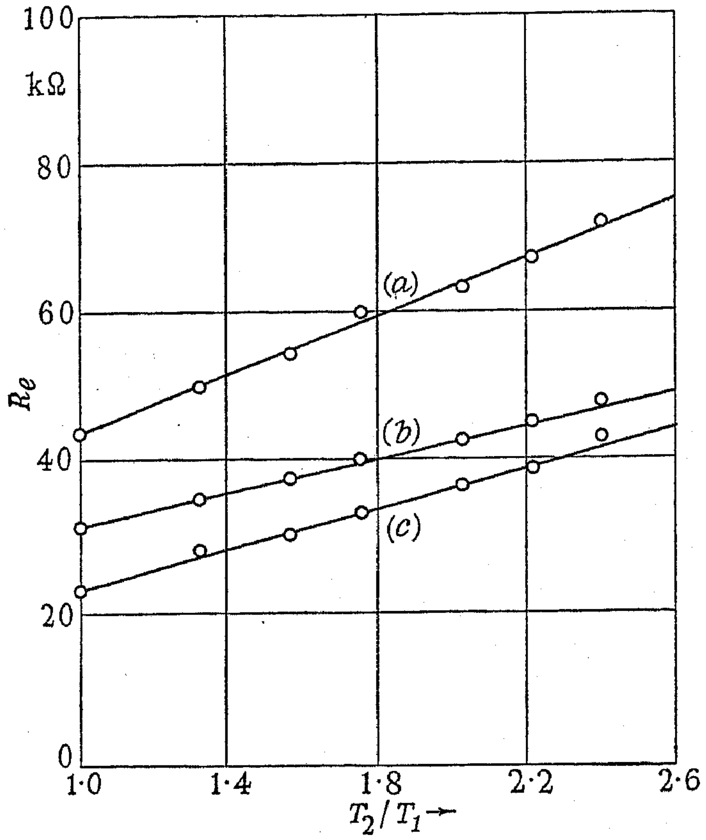

Fig. 9.—As Fig. 8, but with R_1 and R_2 interchanged (see Fig. 7).

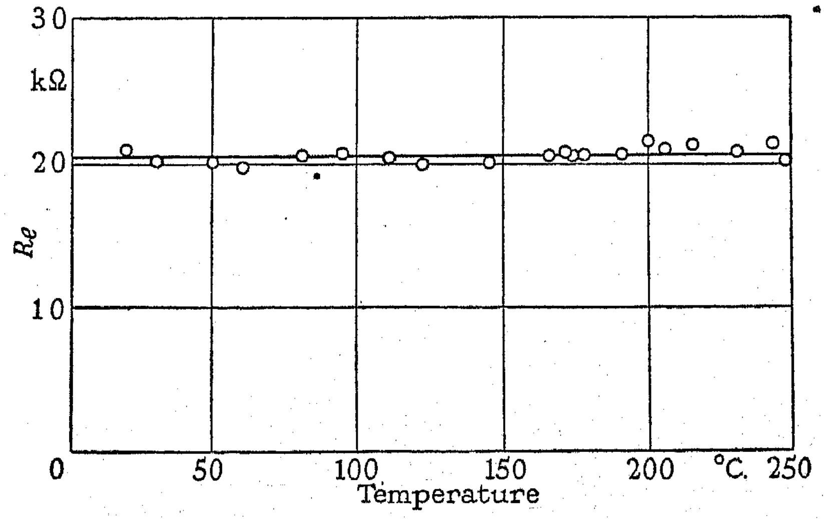

Fig. 10.—Fluctuations in a resonant circuit as a function of the temperature of the condenser.

This result is of much interest when it is remembered that the total fluctuation voltage (i.e. in the frequency range 0 to ∞) of a circuit at uniform temperature is expressed by the equation*

$$\overline{v^2} = \frac{kT}{C}$$

and is independent of R. It follows also that this expression is valid with non-uniform temperature, provided T is the temperature of *all* resistive elements. [Note:—It is now clear that the mere form of equation (2) is insufficient evidence that the fluctuation can be ascribed to R alone, a fact which this last experiment has finally established.]

(6) COEXISTENT THERMAL AND THERMIONIC FLUCTUATIONS

In most analyses of circuit fluctuations the preponderant contribution to the fluctuation can be ascribed to either thermal or thermionic effects: when both are present to a comparable degree the evaluation of the resulting fluctuation presents considerable difficulty. Thus, referring to Fig. 11, there will exist across R a thermal fluctuation due to Brownian motion of the electrons and a further fluctuation due to the passage of the thermionic current. If the current is of suitable value [of the order of $2kT/(eR)$, where e is the electron charge] the magnitudes are comparable. Such low currents must be used if recent advances in the theory of the shot effect are to be investigated experimentally.*

If the current is temperature-limited the two effects are evidently independent, and therefore additive, hence we may write

$$\overline{v^2} = R^2\left(\frac{4kT}{R} + 2Ie\right)df\dagger \quad . \quad . \quad . \quad . \quad (6)$$

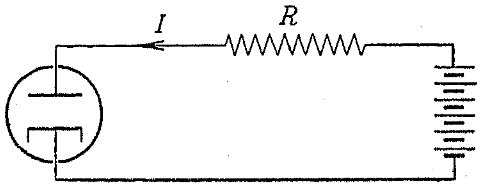

Fig. 11

as the total fluctuation across R in the range df, where e is the electron charge and I is the steady value of the temperature-limited current.

If the current I is space-charge-limited the valve will exhibit a finite slope resistance ρ. Two difficulties are at once encountered. Firstly, fluctuations of anode potential due to thermal agitation in R may dictate the release of electrons from the potential barrier inside the valve, thus vitiating the condition of mutual independence of the fluctuations. Secondly, the mechanism of the addition of the effects is rather obscure. It is possible to argue that mutual dependence of the thermal and thermionic effects will be small since the fluctuations of anode potential are small. This paper will proceed on the assumption that the effects are completely independent, the suitability of the assumption being later verified by a rigorous experimental investigation. The mechanism of the addition has been briefly outlined‡ by the author in an earlier paper, the resulting expression being

$$\overline{v^2} = \left(\frac{R\rho}{R+\rho}\right)^2\left\{\frac{4kT}{R} + 2AIe\right\}df\S \quad . \quad . \quad (7)$$

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* See, for example, Bibliography (3).

* See Bibliography, (7), (8), and (9).
† The second term, $2IeR^2df$, measures the shot contribution.
‡ See Bibliography, (6), Section (3).
§ A is an arbitrary constant fully defined in item (6) of the Bibliography.

This equation can usefully be interpreted with reference to the equivalent circuit of Fig. 11 shown in Fig. 12, in which the valve is represented by the resistance ρ.

Here the thermal fluctuation in R is represented by a hypothetical generator as already described. An analogous generator is shown in series with ρ: this generator must yield an output $\overline{v_s^2}$ such that it represents the thermionic fluctuations generated in the valve. Comparison of (7) with (3), which relates to an identical circuit, shows that the relevant value of $\overline{v_s^2}$ is

$$\overline{v_s^2} = 2AIe\rho^2 df^* \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

This representation of the source of thermionic fluctuations as concentrated in a hypothetical generator has even less physical significance than the similar representation adopted for thermal fluctuations: its adoption, however, permits of the formulation of general equations which simplify the computation of fluctuations in complicated circuits. Thus in any network containing thermionic valves which are replaceable by equivalent resistances (that is, valves operating in the so-called "linear" regime†), the fluctuation appearing between any two points B and B′ can be shown by the process used in deriving equation (1) to be

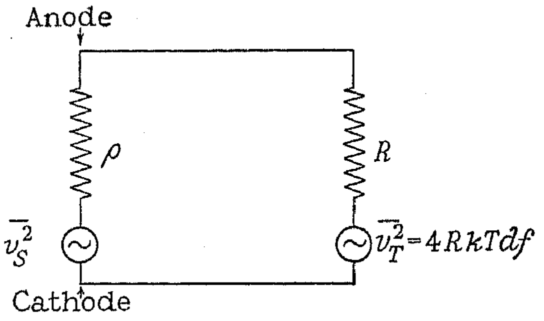

Fig. 12

$$\overline{v^2} = Z_{BB'}^2\left\{4k\sum\frac{R_xT_x}{Z_{Bx}^2} + 2e\sum\frac{A_yI_y}{Z_{By}^2}{\rho_y}^2\right\}df \quad . \quad (9)$$

where A_y, I_y, and ρ_y, are the values of A, I, and ρ, relevant to the typical valve denoted by the subscript y: the remainder of the notation is that of equation (1).

If any of the valves are triodes or multi-electrode valves normally represented as resistances in circuit calculations, there may exist networks in the grid circuits which generate fluctuations. In such cases the grid fluctuation must be separately calculated, multiplied by μ^2 (μ = amplification factor of valve), and added to the fluctuation given by (8). This process follows the usual representation of grid voltages: it is valid only when the grid and anode fluctuations are mutually independent, and the analysis does not therefore apply to retroactive circuits.

The total output of the hypothetical generator in series with ρ is then

$$\begin{aligned}\overline{v^2} &= 2AIe\rho^2 df + \mu^2\overline{v_g^2}df \\ &= (2AIe + g^2\overline{v_g^2})\rho^2 df \quad . \quad . \quad . \quad . \quad . \quad (10)\end{aligned}$$

where $\overline{v_g^2}df$ is the grid fluctuation in the range df, and g is the mutual conductance of the valve. The final representation is shown in Fig. 13. The modification of (9) to allow for $\overline{v^2}_j df$ is self-evident and will not be discussed.

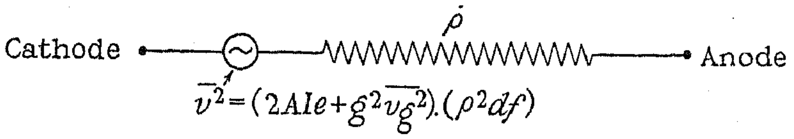

Fig. 13

It may be noted that ρ is often sensibly infinite: no difficulty is experienced, however, provided ρ is first treated as finite and only put equal to infinity when the expression has been cast into a suitable form.

(7) EXPERIMENTAL INVESTIGATIONS OF CO-EXISTENT THERMAL AND THERMIONIC FLUCTUATIONS

The results of Section (6) depend entirely on the assumption of mutual independence of thermal and thermionic effects: the present Section describes experiments designed to test this hypothesis. The first experiments relate to a test of equations (6) and (7); the circuit used is shown in Fig. 14.

R_2 and C are decoupling components, the impedance of C being negligible compared with R at all frequencies to which the amplifier is sensitive. The method of evaluating the fluctuation across R in terms of that across the metallic resistance R' was precisely that described in Section (3). All the apparatus except R remained at T_1 (it was situated in the junction box); R was heated in the special container.

It follows from (7) that the fluctuation across R expressed as equivalent resistance at temperature T_1 is

$$\begin{aligned}R_e &= R\left(\frac{\rho}{R+\rho}\right)^2\left\{\frac{T_2}{T_1} + \frac{2AIeR}{4kT_1}\right\} \\ &= R\left(\frac{\rho}{R+\rho}\right)^2\left\{\frac{T_2}{T_1} + 20AIR\right\} \quad . \quad . \quad (11)\end{aligned}$$

if $T_1 = 290°$ K. In the experiments T_1 varied slightly: a correction—always less than 1 %—was added to the results to compensate for the observed variation.

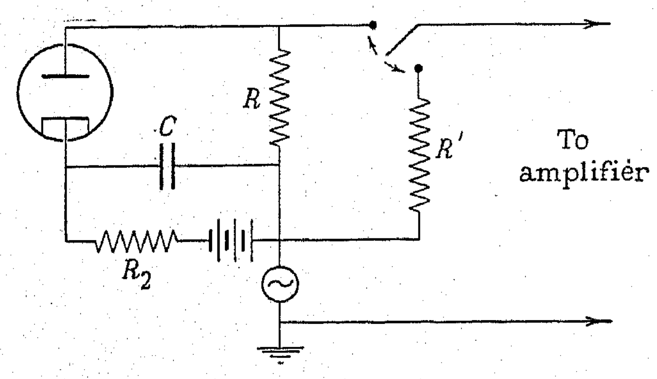

Fig. 14

Fig. 15 shows the results of the test. Curve (a) relates to a directly heated Osram D.E.H.410* valve passing a temperature-limited current of 16 μA: R had a value of $3 \cdot 54$ kΩ and ρ was sensibly infinite. The amplifier passed a 10 % band centred on 8 kc., flicker effect should therefore be negligible and A may be taken as unity. Line (*a*) is plotted from (11); the plotted points are the experimental results and show a maximum discrepancy of -2 % with a mean discrepancy of 0·8 %. Curve (*b*) relates to the more important case of a space-charge-limited current. The valve was an indirectly heated Mazda V914 double diode, one anode only being used. It was chosen because it yielded a low anode current without resorting to a retarding field. The value of $[\rho/(R+\rho)]^2$ as determined by the direct method explained in an earlier paper* was 0·423. R was again 3·54 kΩ. The value of A was deduced from the observation at $T_2/T_1 = 1$; it was 0·665. Agreement is again good, the maximum error being $+2·2$ % and the mean $+0·7$ %.

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* This formula can also be deduced from Pearson's representation of thermionic fluctuations in terms of a resistance $\left(= \frac{1}{2} \cdot \frac{AI}{g^2} \cdot \frac{e}{kT}\right)$ connected in the grid circuit of a triode [see Bibliography, (5)].

† With non-linear operation fluctuations are no longer additive [see Bibliography, (10)].

* Oxide-coated filament.

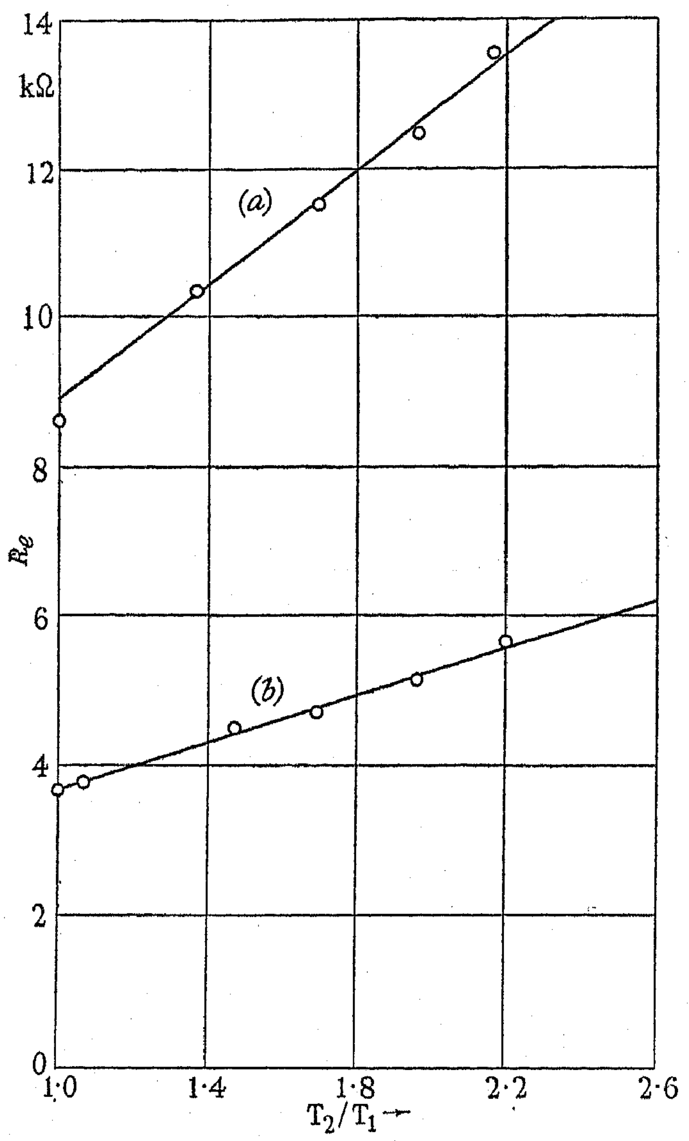

Fig. 15.—Thermal and thermionic fluctuations in the circuit of Fig. 14 as a function of the temperature of R.

Curve (*a*).—Temperature-limited current in Osram D.E.H. 410 valve, oxide-coated filament.

Curve (*b*).—Space-charge-limited current in Mazda V914 valve, oxide-coated cathode.

The straight lines show predicted values. The amplifier passed a 10 % band about a frequency of 8 kc.

Some part of this discrepancy may be due to error in the evaluation of A. Perhaps a preferable value of A would be the mean of the values determined from all the points. This value has been taken in curve (*c*) of Fig. 16, which relates to a similar experiment using a different valve of the same type. In this case the anode current was 20 μA, and $[\rho(R+\rho)]^2$ was 0·435. The deduced value of A varied between 0·89 and 0·81 and had a mean value of 0·865. All points lie close to the calculated line. Curve (*a*) corresponds to (*a*) in Fig. 15, except that the valve was an Osram D.E.8.L.F. with a directly heated thoriated tungsten filament: the current was 10·6 μA, temperature-limited.

Curve (*b*) shows the expectation when both the above currents flow simultaneously through the anode resistance R: this is a final test of the mutual independence of three fluctuation sources. It follows at once from equation (9) and the process used in deriving equation (11) that the expected fluctuation, expressed in terms of equivalent resistance, is given by

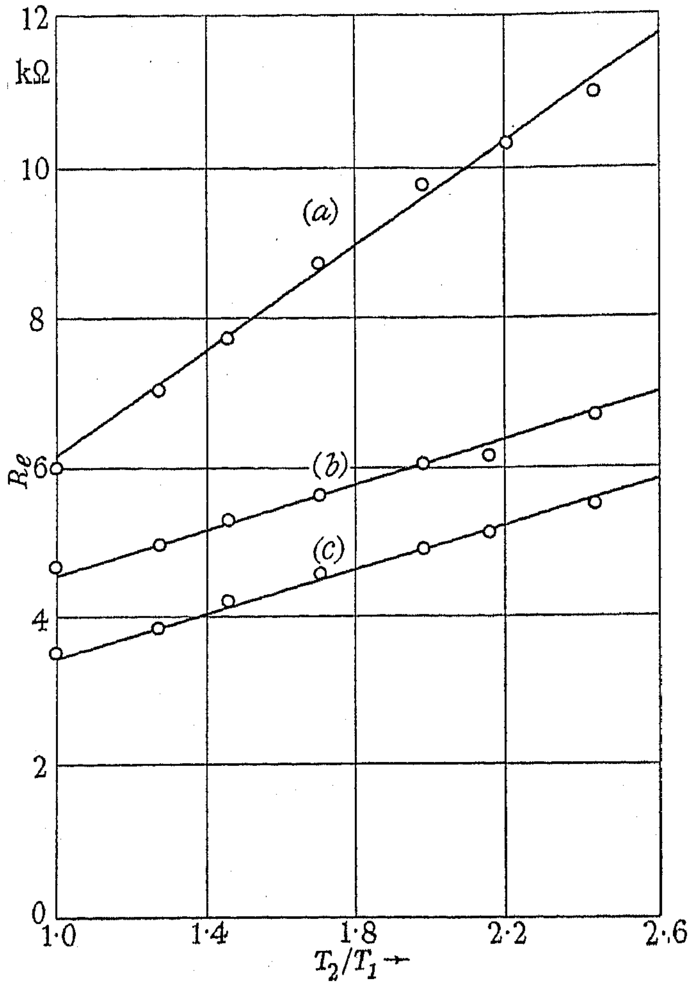

Fig. 16.—As Fig. 15, but with different valves.

Curve (*a*).—Temperature-limited current in Osram D.E. 8 valve, thoriated filament.

Curve (*c*).—Space-charge-limited current in Mazda V914 valve, oxide-coated cathode.

Curve (*b*).—The above temperature- and space-charge-limited currents flowing simultaneously.

The straight lines are predicted values.

$$R_e = R\left(\frac{\rho}{R+\rho}\right)^2\left\{\frac{T_2}{T_1} + 20R(AI_S + I_T)\right\}$$

where I_S and I_T are the space-charge-limited and temperature-limited currents respectively. The straight line (*b*) has been calculated from this equation, using the data relevant to curves (*a*) and (*c*); the small corrections for variation of T_1 were again applied to the experimental results.

The experimental points show a maximum discrepancy of $-3·5$ % and a mean discrepancy of $+0·5$ %.

These experiments are felt to establish the assumed independence of thermal and thermionic effects beyond reasonable doubt: further, they augment earlier evidence* in showing that the equation

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* See Bibliography, (11), Section (2).

$$\overline{v^2} = 2AIe\left(\frac{R\rho}{R+\rho}\right)^2 df$$

is a suitable representation of thermionic fluctuations, whether in fact they are to be ascribed to a thermal or a shot mechanism. They also show that, when this representation is adopted, ρ can safely be regarded as free from fluctuations in value, and that ρ is not then to be regarded as the seat of a coexistent thermal fluctuation, as suggested by Llewellyn.† These results also add to the evidence showing that the doubts very properly expressed by Aldous and Campbell‡ regarding the mathematical adequacy of such a formula as an expression for the fluctuations generated by a space-charge-limited current do not, in practice, limit its application.

(8) APPLICATION TO WIDE-BAND PHOTOCELL AMPLIFIERS

Fig. 17 shows a typical circuit for the first stage of a photocell amplifier. C_1 and C_2 represent the inter-electrode and stray capacitances to earth of grid and anode respectively. The values of the coupling condenser and leak will be supposed such that they can be omitted from the analysis. If the leak is comparable with R_1 the relevant value of R_1 in the analysis is R_1 in parallel with the leak. Following the process outlined in Section (6), the grid fluctuation voltage $\overline{v_g^2}df$ is first calculated by consideration of circuit elements to the left of the dotted line. It follows from equation (9) that

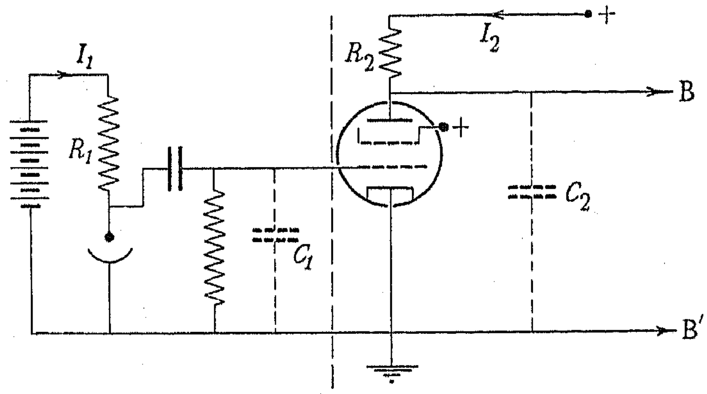

Fig. 17

$$\overline{v_g^2}df = \frac{R_1^2}{1+R_1^2C_1^2\omega^2}\left[\frac{4kT}{R_1} + 2I_1e\right]df \qquad . \quad (12)$$

df being supposed situated about the frequency $\omega/(2\pi)$, and it is supposed that $\rho = \infty$ and $A = 1$ for the photocell.§ The equivalent circuit of the elements to the right of the dotted line is shown in Fig. 18, the fluctuation derived from components to the left being included. Application of formula (9) to this figure yields a fluctuation between B and B′ given by

$$\overline{v^2} = \frac{R^2}{1+R^2C_2^2\omega^2}\left[\frac{4kT}{R_2} + g^2\overline{v_g^2} + 2AI_2e\right]df$$

where $R = R_2\rho/(R_2+\rho)$. In general $\rho >> R_2$, and hence, from (12),

$$\overline{v^2} = \frac{R_2^2}{1+R_2^2C_2^2\omega^2}$$
$$\left[\frac{4kT}{R_2} + \frac{g^2R_1^2}{1+R_1^2C_1^2\omega^2}\left(\frac{4kT}{R_1} + 2I_1e\right) + 2AI_2e\right]df$$

If now a sinusoidal signal of frequency $\omega/(2\pi)$ and mean-square value I_S^2 be superposed on I_1, there will result a signal voltage across R_2 given by

$$V_S^2 = \frac{R_2^2}{1+R_2^2C_2^2\omega^2}\left[\frac{g^2R_1^2}{1+R_1^2C_1^2\omega^2}\right]I_S^2$$

Hence the (noise/signal)² ratio over the band width df is

$$N_\omega^2 = \frac{\overline{v^2}}{V_S^2} = \frac{1}{I_S^2}$$
$$\left[\frac{4kT}{R_1} + 2I_1e + \frac{1+R_1^2C_1^2\omega^2}{g^2R_1^2}\left(2AI_2e + \frac{4kT}{R_2}\right)\right]df$$

which, taking T as 290° K., can be rewritten as

$$N_\omega^2 = \frac{2I_1edf}{I_S^2}$$
$$\left[1 + \frac{0\cdot05}{R_1I_1} + \frac{1+R_1^2C_1^2\omega^2}{g^2R_1^2}\left(\frac{AI_2}{I_1} + \frac{0\cdot05}{I_1R_2}\right)\right] \quad . \quad (13)$$

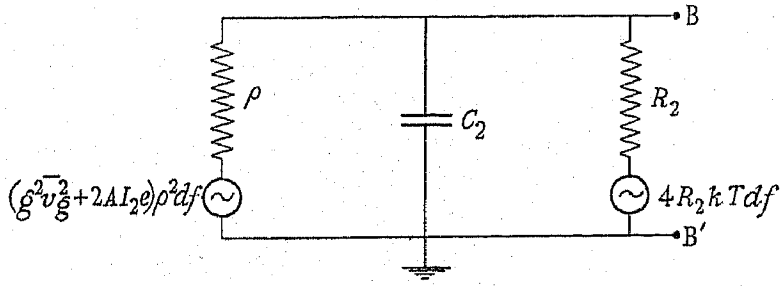

Fig. 18

$2I_1edf/I_S^2$ is the (noise/signal)² ratio in the actual photocurrent and is the lower limit of N_ω^2. The limit is approached only if each of the second two terms in the bracket is much less than 1.

At this point it may be noted that the last term, representing thermal agitation in the anode resistance R_2, can be omitted if $R_2 >> 0\cdot05/(AI_2)$. Since AI_2 will be of the order of 10^{-3}, this condition will always be satisfied.

If now an amplifier is connected across R_2 which has a frequency characteristic such that there results a signal output voltage $v_S = mI_S$, where m is supposed independent of the frequency of I_S over the range $0 - f_{max.}$ and zero outside that range, there will result in the range df an output fluctuation

$$\overline{v^2} = m^2N_\omega^2I_S^2$$

neglecting fluctuations generated in the amplifier itself. The total fluctuation output will be

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* See Bibliography, (6). † *Ibid.*, (4). ‡ *Ibid.*, (14). § *Ibid.*, (13).

$$\overline{v^2} = m^2I_S^2\int_0^{f_{max.}} N^2_\omega$$

yielding an output (noise/signal)^2 ratio

$$N^2 = \int_0^{f_{max.}} N^2_\omega = \frac{2I_1e}{I_S^2}\left[1+\frac{0\cdot05}{R_1I_1}+\frac{AI_2}{I_1g^2R_1^2}\left(1+\frac{R_1^2C_1^2\omega^2{}_{max.}}{3}\right)\right]f_{max.} \quad (14)$$

where $\omega_{max.} = 2\pi f_{max.}$

$2I_1ef_{max.}/I_S^2$ is the (noise/signal)^2 ratio existing in the photo-current, and represents an absolute minimum value for N^2, say N_0^2. Further, N^2 decreases to a limiting value greater than N_0^2 with increasing R_1, for, with R_1 infinite,

$$N^2 = N_0^2\left[1+\frac{AI_2C_1^2\omega^2{}_{max.}}{3I_1g^2}\right] \quad . \quad . \quad . \quad (15)$$

In this expression the second term represents shot effect in the first stage. It follows that the best valve to use for the first stage is that which yields the lowest value for AI_2Cg^2/g^2, where C_g is the grid-earth capacitance of the valve and is supposed a dominant component of C_1. The valve should be operated in a condition giving a minimum value of AI_2/g^2, a condition which has been found before[*] and whose attainment has already been discussed. Values as low as 50 are readily attainable. Hence, taking C_1 as 10 $\mu\mu$F and $\omega_{max.} = 12\cdot56 \times 10^6$,

$$N^2 = N_0^2\left[1+\frac{0\cdot3 \times 10^{-6}}{I_1}\right] \quad . \quad . \quad . \quad (16)$$

Turning now to finite values of R_1, this limit will be approached only if $R_1 >> 1/C_1\omega_{max.}$ and

$$0\cdot05/R_1 << 0\cdot3 \times 10^{-6}, \text{ i.e. } R_1 >> 0\cdot17 \times 10^6$$

Both conditions will be sufficiently satisfied if R_1 is equal to or greater than 1 megohm.

Hence, if R_1 exceeds 1 megohm the (noise/signal)^2 ratio is governed by shot effect of the photo-current and shot effect in the first stage, shot effect in the first stage being dominant or negligible according as I_1 is much greater or much less than $0\cdot3\,\mu A$ $[AI_2C_1^2\omega^2{}_{max.}/(3g^2)$, see equation (15)].

It is believed that television photo-currents are of the order of 10^{-10} amp.; amplifier shot effect is then dominant and the noise/signal ratio is of the order of $55\,N_0$ (see equation 16). If, under similar conditions, R_1 were chosen so as to avoid the necessity for subsequent correction of the frequency characteristic, say $R_1 = 1$ kΩ, we have, from (14), $N > 1\,000\,N_0$. The corrected system utilizing high R_1 is therefore extremely advantageous.

It seems impossible to reduce C_1 appreciably below 10 $\mu\mu$F, and values of N approaching N_0 can only be obtained by either increasing the photo-current to about 10^{-7} amp., or by developing valves for which AI/g^2 is of the order of $0\cdot5$.

It is interesting to note from equation (14) that with sufficiently great photo-currents the signal/noise ratio approximates very closely to the limiting value obtaining in the photo-current. This limit cannot be approached with electron multipliers, for secondary emission will augment the initial fluctuations of the photo-current. It follows that there must exist some value of photo-current below which the electron multiplier is preferable and above which the thermionic amplifier is preferable. This critical value depends on how much secondary emission in the multiplier augments the initial photo-current fluctuations. It seems likely, however, that the critical current will be so high as to lie in a region where background noise is comparatively unimportant.

In the foregoing analysis fluctuations generated in stages subsequent to the first are neglected. Application of the same process to the second stage with the anode current of the first valve replacing the photo-current considered, shows at once that conditions can easily be chosen justifying this procedure. It is necessary only to ensure that the stage gain of the first few stages shall exceed, say, 2 at a frequency of $\omega_{max.}/(2\pi)$. Further, in the integration leading to (14), A was assumed independent of frequency. Such is in fact untrue on account of flicker effect, which increases A at low frequencies. But the influence of flicker effect is confined almost entirely to frequencies less than, say, 2 kc. At these frequencies the third term in equation (13) will be negligibly small compared with its values at higher frequencies, since $R_1 >> 1/(C_1\omega_{max.})$. (In other words equation (13) shows that the relative importance of valve noise in a "corrected" amplifier increases rapidly with the frequency f about which df is situated.) Flicker effect can therefore be neglected and does not invalidate the analysis. It may be noted that this does not hold with "uncorrected" amplifiers, for then $R_1 \simeq 1/C_1\omega_{max.}$ It is therefore a further advantage of the corrected system that flicker effect is effectively suppressed, the fluctuation output of the amplifier due to valve noise being concentrated towards the higher-frequency end of the spectrum.[*] Also grid current has been neglected in the discussion; it will yield an entirely negligible fluctuation provided it is less than, say, 10^{-8} amp., and will not become dominant in the typical case considered till it exceeds $0\cdot34$ μA. An analysis including grid current can readily be developed by the method given above, the grid-current fluctuation being analogous with that produced by the photo-current.

To summarize:—

High values of R_1, say greater than 1 megohm, should be used, the resulting loss of high frequencies being compensated in subsequent stages. The first valve should be chosen and operated such that AIC_g^2/g^2 is a minimum; this is vitally important when the photo-current is less than, say, 10^{-7} amp., for valve noise is then dominant. Grid current should be less than 10^{-8} amp.

Braude[†] has already discussed this problem and has arrived at the conclusion that R_1 should be kept high. He assumed the thermal hypothesis[‡] of valve fluctuations, and accordingly believed that fluctuations generated in screen-grid valves were basically negligible, a belief which has been disproved experimentally.[§]

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[*] See Bibliography (6), Section (5).

[*] Braude's [see Bibliography, (12)] suggested "double amplifier," one dealing with low and one with high frequencies, is therefore unnecessary.

[†] See Bibliography, (12).

[‡] *Ibid.*, (4), (6), and (11).

[§] *Ibid.*, (6).

Further, Braude's analysis omitted the fluctuations generated in the photo-current itself. These two sources of fluctuations have been shown to determine the practical and ideal limits of background noise respectively. Hence Braude's calculation of these limits and of the improvement obtained by increasing R_1 are at fault, as also is his proposed method of eliminating background noise by using reactive impedances in the early stages. With values of R_1 greater than 1 megohm thermal noise is negligible; his proposed values of R_1, between $10^7\,\Omega$ and $10^9\,\Omega$, are therefore unnecessarily high.

(9) CONCLUSIONS

The first series of experiments establishes the general formula (1) relating to thermal fluctuations in complex networks with non-uniform temperature distribution. The thermal effects can be represented satisfactorily by concentrated e.m.f.'s associated solely with the resistive elements of the network: the values of the e.m.f.'s depend only on the values and temperatures of the resistances—the temperatures of reactive elements do not influence thermal fluctuations.

The second series of experiments establishes the mutual independence of thermal and thermionic effects, and proves the suitability of the formula used to represent thermionic fluctuations. The slope resistance of the valve must there be taken as constant, and thermal fluctuations must not be ascribed to it.

In wide-band photocell amplifiers a "corrected" system employing a high resistance of the order of 1 megohm in the anode circuit of the photocell is vastly preferable to an "uncorrected" system with its associated low anode resistance. In typical circumstances thermionic noise in the first amplifying valve will be dominant. Accordingly, the first valve should be chosen to yield the lowest possible value of AIC_1^2/g^2. The operating conditions should be such that AI/g^2 is a minimum, provided g is not thereby reduced to such an extent that the gain of the first stage at the highest wanted frequency falls below 2. In "corrected" amplifiers the noise output is concentrated towards the high-frequency end of the spectrum, and flicker effect can be neglected.

There exists a critical value of photo-current above which the thermionic amplifier yields a better signal/noise ratio than the electron multiplier. This critical current cannot be evaluated without further data relating to electron multipliers; it is probably so high as to lie in a region where background noise is comparatively unimportant.

(10) ACKNOWLEDGMENTS

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DISCUSSION ON
"THERMAL FLUCTUATIONS IN COMPLEX NETWORKS"*

Mr. D. A. Bell (*communicated*): I believe that the analysis given in Appendix 1 of the paper can be made of even greater generality by a very slight change in the handling of the transfer impedances.

Since I have indicated in a recent paper† that the fundamental property of a pure resistance is probably the generation of the fluctuation current

$$\overline{I}^2 = \frac{4kT}{R}df \quad . \quad . \quad . \quad . \quad . \quad . \quad (A)$$

rather than the voltage $\overline{V}^2 = 4RkTdf$, it is interesting to repeat the analysis in terms of a fluctuation current in place of voltage. In order to avoid confusion with the author's notation, while using his Fig. 11, the letter N with suitable suffixes will be used to represent various impedances, defined as follows:—

(1) Let N_{Ax} be equal to the p.d. between the points A and A′ (on open circuit externally) per unit current in r_x.

(2) Let $N_{AA'}$ be the internal impedance between A and A′, i.e. the p.d. between A and A′ per unit current injected from a generator connected between these two points.

(3) Let $1/N_{xA}$ be equal to the current in r_x per unit p.d. across AA′, i.e. N_{xA} is the impedance from A to x corresponding to the impedance N_{Ax} from x to A.

Assuming an input mean square current i_x^2 flowing in r_x, we then have an output mean-square voltage across AA′ given by

$$|v^2_{AA'}| = N^2_{Ax} i^2_x$$

and, using equation (A),

$$|v^2_{AA}| = N^2_{Ax} \cdot \frac{4kT_x}{r_x}df$$

$$\therefore \quad |i^2_{AA'}| = \frac{|v^2_{AA'}|}{N^2_{AA'}} = \frac{4kdf}{N^2_{AA'}} \cdot \frac{N^2_{Ax}T_x}{r_x}$$

For a number of resistive elements r_x within the network, the resultant mean square fluctuation current between A and A′ will be

$$|I^2_{AA'}| = \frac{|V^2_{AA'}|}{N^2_{AA'}} = \frac{4kdf}{N^2_{AA'}} \cdot \sum \frac{N^2_{Ax}T_x}{r_x} \quad . \quad . \quad (B)$$

Now let a current i_0 from an external generator be injected throughout the points A, A′. The mean square current in r_x will then be

$$|i^2_x| = \frac{E^2_{AA'}}{N^2_{xA}} = \frac{i^2_0 N^2_{AA'}}{N^2_{xA}}$$

and the power dissipated therein is

$$|i^2_x| r_x = \frac{r_x i^2_0 N^2_{AA'}}{N^2_{xA'}} \quad . \quad . \quad . \quad . \quad (C)$$

But if the input impedance $N_{AA'}$ as viewed from the external generator is made up of resistive and reactive components in series, $m + jn$, the power absorbed from the generator will be $i^2_0 m$. Comparing this with the power dissipation given by equation (C), we find that

$$m = \frac{r_x N^2_{AA'}}{N^2_{xA}}$$

or, in the general case of many resistive elements,

$$|m| = N^2_{AA'} \sum \frac{r_x}{N^2_{xA}} \quad . \quad . \quad . \quad . \quad (D)$$

Comparing equations (B) and (D), we find that, provided $N^2_{Ax} = N^2_{xA}$,

$$|I^2_{AA'}| = 4kdf \sum \frac{T_x}{m_x}$$

corresponding exactly to the result obtained by the author in the analysis based on fluctuation voltage.

Now he demanded, as necessary for the validity of his analysis, the condition $Z_{Ax} = Z_{xA}$, and therefore limited himself to linear networks. But, as presented above (and the same could be shown to be true of the author's analysis), the necessary condition is a relationship between squares of impedances, $N^2_{Ax} = N^2_{xA}$, and it appears that this must be satisfied even by non-linear networks. For consider the system shown in Fig. A, where two equal resistances R at the same temperature T are joined respectively to pairs of terminals 1 and 2 of a non-linear impedance supposed to be such that $Z_{12} \neq Z_{21}$, but both Z_{12} and Z_{21} are large compared with R. Now, on any hypothesis, two equal resistances at the same temperature will generate equal apparent fluctuation voltages, say of mean-square value V^2. Then, since Z is large compared with R, the two currents flowing in the circuit are approximately

$$I^2_{12} = V^2/Z^2_{12}$$

$$I^2_{21} = V^2/Z^2_{21}$$

The amounts of energy transferred from resistance 1 to resistance 2 and from resistance 2 to resistance 1 are then

$$\left.\begin{aligned} W_{12} &= RI^2_{12} = RV^2/Z^2_{12} \\ W_{21} &= RI^2_{21} = RV^2/Z^2_{21} \end{aligned}\right\} \quad . \quad . \quad . \quad (E)$$

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* Paper by Dr. F. C. WILLIAMS (see page 53). Reprinted from *Journal I.E.E.*, 1938, vol. 83, p. 432.

† "A Theory of Fluctuation Noise," *Journal I.E.E.*, 1938, vol. 82, p. 522; and *Proceedings of the Wireless Section*, 1938, vol. 13, p. 97.

But unless $W_{21} = W_{12}$ the resistance receiving the greater amount of energy will rise in temperature, and heat will then be transferred in a self-acting system from a body at a lower temperature to one at a higher temperature. This, however, is contrary to the second law of thermodynamics* and is thus impossible. Therefore, from the necessity of equating W_{12} and W_{21} in equations (E), we are compelled to put

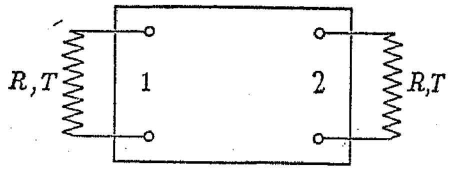

Fig. A

$$Z_{12}^2 = Z_{21}^2$$

It follows that the theorem proved by the author, that the fluctuation voltage at the terminals of a complex network is of magnitude corresponding to the resistive component of the impedance as viewed from the terminals, is also true of non-linear systems, provided only that the resistive component of the impedance is calculated in terms of power dissipation. I have already discussed the application of this principle to the thermionic valve (in the paper quoted above); but is it not also possible that the iron-cored coil, for which experimental figures are given in Fig. 8 of the present paper, was in fact a non-linear system if hysteresis losses were appreciable? It would be interesting to try the experiment on a more thoroughly non-linear device, e.g. an iron-cored coil without air-gap, and perhaps with a suitable polarizing current. In such cases, where it is necessary to discriminate against harmonics, the resistive component of the input impedance could probably be measured better by a milliammeter and dynamometer method than by the three-voltmeter method.

Dr. F. C. Williams (*in reply*): Mr. Bell's proposal of a theorem in terms of a "current representation" of fluctuations comparable with that given in the paper in terms of a "voltage representation" is of considerable interest. The idea is not new,† but has never been fully discussed. I cannot agree, however, that Mr. Bell's analysis is equivalent to that developed in the paper. His analysis leading to equation (D) is a repetition of Appendix 2, and comparison of the notations in the resulting equations shows that

$$|m| \equiv r; \quad N_{AA'}^2 \equiv Z_{AA'}^2; \quad N_{Ax}^2 \equiv Z_{Ax}^2$$

From Mr. Bell's equation (B),

$$\bar{V}_{AA'}^2 = 4kdf \sum \frac{N_{Ax}^2 T_x}{r_x}$$

But, expressing equation (2) of the paper in Mr. Bell's notation by means of the identities given above,

$$\bar{V}_{AA'}^2 = N_{AA'}^2 4kdf \sum \frac{r_x T_x}{N_{xA}^2}$$

These two expressions for $\bar{V}_{AA'}^2$ are not equivalent.*

The discrepancy is thought to be due to a fundamental error in Mr. Bell's interpretation of equation (A). He states that a resistance R generates a current of mean square value defined by equation (A), and it appears from his definition of N_{Ax} that this current is supposed to flow always in R, whether R is connected to a network or not; he apparently identifies $\bar{I}^2$ with the thermal motion of free electrons in R. In my view, equation (A) defines a current supplied to R by a hypothetical generator connected in parallel with R. In such circumstances $\bar{I}^2$ flows partly in R and partly in the network to which R is connected. Also, $\bar{I}^2$ then bears no relation to the actual motion of free electrons and is merely a mathematical tool which assists computation; it is no more fundamental than is the "voltage representation" given in the paper. These views are fully set out in a paper at present in course of preparation, and need not be further discussed here, except to note that a "current representation" based on them has been developed which yields results consistent with those obtained by "voltage representation."

The limitation of the analysis to linear networks does not depend only on the requirement that $Z_{Ax}^2 = Z_{xA}^2$, for the summation leading to equation (2) assumes that the mean square value of the sum of a number of fluctation currents is the sum of their mean-square values, and such is not necessarily true in non-linear networks. Further, it is not usually possible to specify impedance values in non-linear networks: the "impedance" then depends in general on the amplitude and wave-form of the applied e.m.f. and on the presence or otherwise of other e.m.f.'s in the network. Thus, despite Mr. Bell's ingenious proof that Z_{12}^2 is always equal to Z_{21}^2 (since both can vary with time in non-linear networks, this should perhaps be stated: average Z_{12}^2 = average Z_{21}^2), it is still thought necessary to restrict the analysis to linear networks. The limitation of scope involved appears small, since there are but few common circuit-elements which respond in a markedly non-linear manner to the exceedingly small currents and potentials associated with fluctuation phenomena.

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* That is to say, it would be a system in which "perpetual motion" could be obtained by utilizing the self-maintaining difference of temperature to drive a heat engine.

† *Journal I.E.E.*, 1936, vol. 79, p. 349; and *Proceedings of the Wireless Section*, 1936, vol. 11, p. 316.

* It can be shown that, as defined by Bell, N_{xA} is not equal to N_{Ax}.

MEETINGS OF THE WIRELESS SECTION

142nd MEETING OF THE WIRELESS SECTION, 2nd MARCH, 1938

Mr. T. Wadsworth, M.Sc., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting held on the 2nd February, 1938, were taken as read and were confirmed and signed.

A paper by Mr. P. P. Eckersley, Member, entitled "A Quantitative Study of Asymmetric-Sideband Broadcasting" (see page 137), was read and discussed.

A vote of thanks to the author, moved by the Chairman, was carried with acclamation.

13th INFORMAL MEETING OF THE WIRELESS SECTION, 15th MARCH, 1938

Mr. T. Wadsworth, M.Sc., Chairman of the Section, took the chair at 6.30 p.m.

The minutes of the Informal Meeting held on the 25th January, 1938, were taken as read and were confirmed and signed.

A discussion, opened by Mr. R. C. G. Williams, Ph.D., Graduate, took place on "The Servicing of Broadcast Receivers."

At the conclusion of the discussion a vote of thanks was accorded to Mr. Williams for his introductory remarks.

143rd MEETING OF THE WIRELESS SECTION, 30th MARCH, 1938

Mr. T. Wadsworth, M.Sc., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting held on the 2nd March, 1938, were taken as read and were confirmed and signed.

A paper by Messrs. J. Bell, B.Sc., J. W. Davies, and B. S. Gossling, M.A., entitled "High-Power Valves: Construction, Testing, and Operation" (see page 177), was read and discussed.

A vote of thanks to the authors, moved by the Chairman, was carried with acclamation.

144th MEETING OF THE WIRELESS SECTION, 6th APRIL, 1938

Mr. T. Wadsworth, M.Sc., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting held on the 30th March, 1938, were taken as read and were confirmed and signed.

A paper by Messrs. A. J. Gill, B.Sc.(Eng.), Member, and S. Whitehead, M.A., Ph.D., Associate Member, entitled "Electrical Interference with Radio Reception" (see page 209), was read and discussed.

A vote of thanks to the authors, moved by the Chairman, was carried with acclamation.

145th MEETING OF THE WIRELESS SECTION, 4th MAY, 1938

Mr. W. J. Picken, Vice-Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting held on the 6th April, 1938, were taken as read and were confirmed and signed.

The Chairman announced that the following members had been nominated to fill the vacancies which would occur on the Committee on the 30th September, 1938:—

Chairman: A. J. Gill, B.Sc.(Eng.).

Vice-Chairman: H. Bishop, B.Sc.(Eng.).

Ordinary Members of Committee: A. J. A. Gracie, B.Sc., G. S. C. Lucas, J. S. McPetrie, B.Sc., Ph.D., Col. G. D. Ozanne, M.C., W. J. Picken, and R. T. B. Wynn, M.A.

In the event of a ballot for the new Committee being required, Messrs. F. Jervis Smith and J. P. E. Vigoureux were appointed scrutineers.

Dr. Ralph Bown, M.E., then delivered a lecture on "Researches in Radiotelephony" (see page 259).

A vote of thanks to the lecturer, proposed by Mr. H. L. Kirke and seconded by Mr. A. J. Gill, was carried with acclamation.